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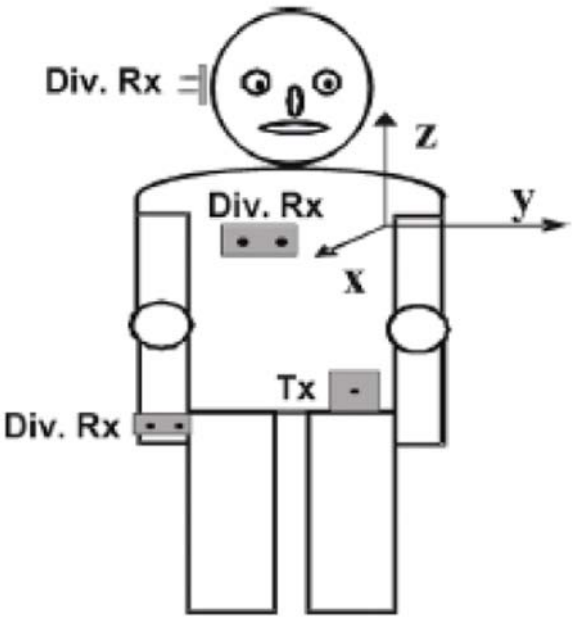

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St.-Pietersnieuwstraat 41, B-9000 Gent (Belgium)

Contents

Front cover: "The locations of a transmitter and three space-diversity receivers on a human body, used for experiments in characterizing wireless channels on the human body. See the paper by Peter S. Hall on pp. 12-25.

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Editorial

The Deadline is Approaching!

This is the last issue of the *Radio Science Bulletin* before the deadline for submission of papers to the XXX General Assembly and Scientific Symposium of the International Union of Radio Science. The deadline is **February 11, 2011**. The General Assembly and Scientific Symposium (GASS) will be held August 13-20, 2011, in Istanbul, Turkey. The call for papers appears in this issue. There are also announcements of how to apply for the URSI Young Scientist Awards, and how to enter the URSI Student Paper Competition. You need to take action on this now.

In addition to being a great scientific meeting, it is significant that the GASS is being held in Istanbul. I have had the pleasure of visiting Istanbul several times in the past, and I'm very much looking forward to visiting it again. It is beautiful, modern city with wonderful people and a tremendously rich and fascinating history. The historic sights are not to be missed. If at all possible, plan on bringing your family: this is a *very* special place!

Our Papers

There are a growing number of applications in which receivers and transmitters are mounted on the human body and communicate with each other. Examples include medical sensing and monitoring, monitoring human performance during sporting events, entertainment, and communications. The propagation channel on the human body becomes quite important in such applications. Characterizing this channel is the topic of the invited *Review of Radio Science* paper by Peter Hall and Yang Hao. The authors begin by reviewing the applications and frequency ranges used by on-body communications. Because the communication systems can use either narrowband or wideband communications, the characteristics of the on-body channel for each are treated separately. The authors explore the models that have been developed for the variation in path gain with distance. They then present the statistics of the channel, and explain how these vary depending on the motion of the body. These results are based on an extensive series of measurements, made using male and female subjects wearing on-body transceivers with a variety of antennas. The path-loss and time-delay effects are then presented for the wideband on-body channel. The results of experiments with multiple-antenna systems are presented. This includes the use of multiple antennas in diversity systems, as well their use in multiple-input multiple-output (MIMO) systems.

The efforts of Giuliano Manara and Phil Wilkinson in bringing us this *Review* are gratefully acknowledged.

The estimation of the frequency and the phase of a received signal when the complex amplitude of the signal is non-constant and unknown is the topic of the invited *Review of Radio Science* paper by N. Noels, P. Ciblat, and H. Steendam. This is an important problem in many areas of communications. The authors begin by explaining the problem and providing the motivation for its solution. Communications over a noisy channel usually require that the transmitter and receiver have carriers with nearly the same frequency and phase. Drifts in the oscillators, channel effects, noise, and Doppler shifts can cause variations in the frequency and phase. The receiver estimates these variations, and corrects the frequency and phase of the carrier as part of the detection process. The accuracy with which this estimation can be done therefore has a direct effect on the accuracy of the detection. The authors explain that there are several different lower bounds on this estimation accuracy. Among these are the Cramer-Rao bound, a modified Cramer-Rao bound, the Barankin bound, and the Ziv-Zakai bound. The authors present and explain the mathematical model used for analyzing these various bounds. They then derive the Cramer-Rao bound. They examine its use in different situations, including where the non-constant complex amplitude of the signal is due to digital modulation or is due to a Gaussian process. They look at the values of the bound in various asymptotic cases. They then repeat these analyses for the modified Cramer-Rao bound, the Barankin bound, and the Ziv-Zakai bound. The results of a series of simulations are presented, which serve to illustrate the performance of these different bounds for various situations associated with digital communications. These results lead to conclusions regarding which bound provides the best estimate of the performance that can be achieved for the various situations.

The efforts of Marco Luise and Phil Wilkinson in providing this *Review* are appreciated.

The URSI Working Group on Natural and Human Induced Hazards (WG1) was created at the 2008 General Assembly in Chicago. The topic of this Working Group is of tremendous importance: it deals with how radio science can help to save lives. In their paper, P. J. Wilkinson and D. G. Cole introduce the Working Group, and explain its goals and efforts to date. They review the stages in the reduction of risks from hazards, and how radio science can play a role. This includes predicting and preventing disasters, reducing disaster risks, immediate disaster response, supporting disaster relief, and learning from disasters. The authors look at the possible roles of radio science in disaster management. Areas where radio science could contribute include communications, propagation, data mapping and analysis, remote sensing, signal processing, and the development of information packages. While this is a report on the results from the Working Group to date, it is also a call for input. This vitally important topic reaches across almost all Commissions and areas of interest in radio science. *Your input can help*. Please read the paper, and then respond to Phil Wilkinson with your ideas.

Also in this Issue

It is with great sadness that I report the loss of Carl Baum and Paul Kintner. They are remembered in this issue.

Kristian Schlegel has again provided us with a review of an interesting new book. If you have a book you think should be reviewed, please contact Kristian.

I'll end my comments where I began: send in your paper(s) for the URSI General Assembly and Scientific Symposium in Istanbul! The Web site is now open for submissions: http://www.ursigass2011.org. You really will want to be a part of what is going to be an outstanding meeting.

My very best wishes to all our URSI family for a most joyous holiday season, and for a very happy, healthy, safe, and prosperous New Year.

Ross

XXV IUGG General Assembly

Melbourne Convention and Exhibition Centre, Australia, June 28 - July 7, 2011

The *International Union of Geodesy and Geophysics* (IUGG) 2011 Scientific Program Committee invites the submission of abstracts on original work to be considered for oral or poster presentation at the IUGG 2011 General Assembly. Information about the program, registration, accommodation, and sponsorship can be found at

http://www.iugg2011.com.

Key Dates

Call for abstracts closes (for grant applicants only):
January 17, 2011

Call for abstracts closes (for all other submissions):
February 1, 2011

Authors and grant applicants notified of acceptance:
March 28, 2011

Author/grant applicants registration and early bird deadline: April 11, 2011

IN MEMORIAM

Carl Edward BAUM
1940 - 2010

His ideas kept flowing like a mighty river

Carl Baum, mentor to many, took his last breath peacefully on December 2, 2010, in Albuquerque, New Mexico. Carl was born in Binghamton, New York, on February 6, 1940. He received his BS (with honors), MS, and PhD degrees in Electrical Engineering from the California Institute of Technology, Pasadena, California, in 1962, 1963, and 1969, respectively. Following his BS, he received his commission in the Air Force, and was stationed at the Air Force Weapons Laboratory at Kirtland AFB, Albuquerque, New Mexico. He served from 1963 until 1971 as an officer, and then accepted a civilian position, and retired as a Senior Scientist in 2005. Since his retirement from USAF, he was a Distinguished Professor in the Department of Electrical and Computer Engineering at the University of New Mexico.

During his military career, he was awarded the Air Force Research and Development Award and the Air Force Nomination to Ten Outstanding Young Men of America. In a career that spanned five decades, this remarkably creative engineer introduced innumerable new concepts in mathematics, electromagnetic theory, and system design, many of which remain the standards of excellence today. From his earliest designs in EMP sensors and simulators, to the latest developments in high-power microwave (HPM) and ultra-wideband antenna and system design, Dr. Carl Baum's research has remained ever on the forefront of technology. His advances in EM theory have left an indelible mark and a lasting legacy on the technical world, and have led to much of what we do today in EMP, HPM, and target identification.

His scientific contributions were prodigious. He wrote innumerable technical notes, articles, books, and presentations. He was the editor of the *Note Series* that has published state-of-the-art research results for the past 45 years. He received the Richard R. Stoddart award of the IEEE EMC Society (1984), the Harry Diamond Memorial Award (1987), the AFSC Harold Brown Award (1990), and the Air Force Basic Research Award (Honorable Mention) in 1999. In addition, he received five Best Paper Awards from the AMEREM/ EUROEM Awards Committee. He and his research team were honored as an AFOSR Star Team for 2000-2002, and received the first annual R. Earl Good Award from AFRL (2004) for their work in target identification. He was named an IEEE Fellow in 1984, an EMP Fellow in 1986, and the first Air Force Research Laboratory Fellow in 1996. However, the honors that meant the most to him came in July of 2004, when he was bestowed with an Honorary Doctorate of Engineering by Otto von Guericke University in Magdeburg, Germany, during EUROEM 2004, and received a special honor from his colleagues in Russia for his lifetime of achievements. He received the IEEE John Kraus Antenna Award (2006) and the IEEE Electromagnetics Award (2007). He was a member of Commissions A, B, and E of the US National Committee of the International Union of Radio Science (URSI). He established the SUMMA Foundation, which sponsors various electromagnetics-related activities, including scientific conferences, publications, short courses, fellowships, and awards. He led EMP short courses and high-power electromagnetics workshops around the globe. Dr. Baum was an active organizer of scientific conferences and workshops that brought together researchers from all over the United States and the world to share the latest in electromagnetic research.

When not putting his new ideas in mathematics and electromagnetics into new technical notes or organizing meetings, Dr. Baum enjoyed playing the piano and creating his own musical compositions. Many of these have been heard at the biennial AMEREM and EUROEM conferences. His compositions can also be heard at one of the many churches in Albuquerque that host the annual concerts of the Albuquerque Symphony Orchestra and Chorus, and even at his own church, where he used to be the choir director. Twenty-three of these compositions have been recorded.

Carl is survived by his two nephews, George and Spenser, and his sister-in-law, Martha Baum, all of Albuquerque.

Provided by David V. Giri (e-mail: Giri@DVGiri.com), with input from many of Carl's friends.

Paul M. Kintner, Jr.
1946 - 2010

Paul M. Kintner, Jr., Professor of Electrical and Computer Engineering and Head of the Global Positioning Systems Laboratory at Cornell University, died at his home in Ithaca on Tuesday, November 16, 2010, after a courageous battle with pancreatic cancer. Kintner was an internationally recognized authority on the interaction of radio signals – both natural and man-made – with space environments, particularly with the ionosphere and magnetosphere. His studies included the effect of the space environment on GPS signals. During the 2009-10 academic year, he served as a Jefferson Science Fellow at the US Department of State, advising the government on GPS, navigational satellite systems, space weather, and other scientific topics with implications for defense and national security.

Prof. Kintner earned his undergraduate degree from the University of Rochester in 1968. He received a PhD in Plasma Physics from the University of Minnesota in 1974 for work on the space environment of the northern lights. He continued this work with the Space Physics Group at the University of Iowa until 1976, when he moved to Cornell as a research associate. He was appointed to the Cornell faculty in 1981.

He was a Fellow of the American Physical Society, and a Senior Member of the Institute of Electrical and Electronics Engineers. He chaired the Living with a Star/Geospace Mission Definition Team, and NASA's Sun-Earth Connections Advisory Subcommittee. He served on National Research Council committees on "Solar and Space Physics" and on the "Economic and Societal Impacts of Severe Solar Storms." In 2007, he convened an American Geophysical Union Chapman Conference on "Mid-Latitude Ionospheric Dynamics and Disturbances," leading to a monograph by the same name. In September 2009, he delivered the Birkland Lecture to the Norwegian Academy of Science and Letters.

He was a mentor to generations of Cornell students and younger faculty members, often at pivotal points in their professional development. He continued to advise graduate students and colleagues preparing for a sounding-rocket research campaign in Norway until days before his death.

He is survived by his wife, Constance Bart Kintner, with whom he shared the love of family and wilderness pursuits. His also survived by his children: Douglas T. S. Kintner of Ithaca, New York; Paul M. S. Kintner, at the University of Rochester; Robert Bart of Hood River, Oregon; Rebecca Bart of Berkeley, California; his son-in-law, Kater Murch, and his grandson, West Bart Murch, also of Berkeley, California. Also surviving are his father, Dr. Paul M. Kintner, Sr., and his mother, Vivian Kintner, of Hendersonville, North Carolina; brothers Douglas Kintner of Sun Prairie, Wisconsin, and Christopher Kintner of Del Mar, California; sister Victoria Kintner Griswold of Indianapolis, Indiana; and several aunts, uncles, nieces, and nephews, many of whom he was able to visit in his final months. He was predeceased by his first wife, Janet Rae Smith-Kintner.

Donations in his memory can be made to the Department of Electrical and Computer Engineering at Cornell University, or to Hospicare and Palliative Care Services of Tompkins County.

[The photo is courtesy of the Division of Rare and Manuscript Collections, Cornell University Libraries. The text was provided by the family of Paul Kintner, with thanks to Laurie Shelton of Cornell and Michael Rietveld of URSI Commission G.]

XXX General Assembly and Scientific Symposium of the International Union of Radio Science

Union Radio Scientifique Internationale

August 13-20, 2011 Lütfi Kırdar Convention and Exhibition Centre, Istanbul, TURKEY

Call for Papers

The XXX General Assembly and Scientific Symposium of the International Union of Radio Science (Union Radio Scientifique Internationale: URSI) will be held at the Lütfi Kırdar Convention and Exhibition Centre in the beautiful historical center of Istanbul, Turkey, August 13-20, 2011.

The XXX General Assembly and Scientific Symposium will have a scientific program organized around the ten Commissions of URSI and consisting of plenary lectures, public lectures, tutorials, posters, invited and contributed papers. In addition, there will be workshops, short courses, special programs for young scientists, student paper competition, programs for accompanying persons, and industrial exhibits. More than 1,500 scientists from more than fifty countries are expected to participate in the Assembly. The detailed program, the link to an electronic submission site, the registration form, and hotel information will be available on the General Assembly Web site: **http://www.ursigass2011.org**

Information for all authors -*Submission information*
All contributions (four pages full paper and up to 100 words abstract) should be submitted electronically via the link provided on the General Assembly Web site. Please consult the symposium Web site, **http://www.ursigass2011.org**, for the latest instructions, templates, and sample formats.

Important Deadlines

Paper submission	February 11, 2011
Notification of acceptance	April 30, 2011

Topics of Interest
Commission A : Electromagnetic Metrology Commission B : Fields and Waves Commission C : Radiocommunication Systems and Signal Processing Commission D : Electronics and Photonics Commission E : Electromagnetic Environments and Interference Commission F : Wave Propagation and Remote Sensing Commission G : Ionospheric Radio and Propagation Commission H : Waves in Plasmas Commission J : Radio Astronomy Commission K : Electromagnetics in Biology and Medicine

Student Paper Competition
A student must be first author of the paper. The student's advisor should attach a statement that his/her contribution is primarily advisory. All other submission requirements and instructions can be found at symposium Web site.

Special Sessions
Individuals interested in organizing special sessions should request permission from the Chair of the appropriate URSI Commission.

Contact
For any questions related to the XXX General Assembly, please contact the Chair of the Conference: Prof. Hamit Serbest Department of Electrical and Electronics Engineering Cukurova University, Adana, Turkey
E-mail: ursigass2011@ursigass2011.org

FIRST ANNOUNCEMENT

The XXX General Assembly and Scientific Symposium of the International Union of Radio Science (Union Radio Scientifique Internationale-URSI) will be held at the Lütfi Kırdar Convention & Exhibition Centre, Istanbul, Turkey on August 13-20, 2011.

The General Assemblies and Scientific Symposia of URSI are held at intervals of

plans for future research and special projects in all areas of radio science, especially where international cooperation is desirable. The first Assembly was held in Brussels, Belgium in 1922 and the latest in Chicago, IL, USA in 2008.

The XXX General Assembly and Scientific Symposium will have a scientific program organized around the ten Commissions of URSI and consisting of plenary lectures, public lectures, tutorials, invited and contributed papers. In addition, there will be workshops, short courses, special programs for young scientists, student paper competition, and programs for accompanying persons. More than 1,500 scientists from more than fifty countries are expected to participate in the Assembly and Scientific Symposium.

The Call for Papers will be issued in mid 2010, will be published in the Radio Science Bulletin and in the IEEE Antennas and Propagation Magazine, and will be posted on the URSI website. It is expected that all papers should be received by the beginning of February 2011, that Authors will be notified of the disposition of their submissions by the end of April 2011.

Preliminary information

Detailed information on the scientific program and on abstract submissions will be available toward the end of March 2010. A web site with current information on the XXX General Assembly is available at: **www.ursigass2011.org** and all abstracts will be received electronically.

Organizing Committee

Chair	A. Hamit Serbest, *Cukurova University, Turkey*, hamitserbest@gmail.com
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Student Paper Competition	Birsen Saka, *Hacettepe University, Turkey*, birsen@hacettepe.edu.tr
Workshops/Short Courses	Vakur B. Ertürk, *Bilkent University, Turkey*, vakur@ee.bilkent.edu.tr
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Social activities	Tour Select Ltd, organization@ursigass2011.org
Hotel Reservation	Tour Select Ltd, organization@ursigass2011.org
Contact with URSI	P. Van Daele, peter.vandaele@intec.ugent.be W. Ross Stone, r.stone@ieee.org

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Coordinator	P.L.E. Uslenghi *(U.S.A)*
Associate Coordinator	A. Hamit Serbest *(Turkey)*
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www.ursigass2011.org

AWARDS FOR YOUNG SCIENTISTS

CONDITIONS

A limited number of awards are available to assist young scientists from both developed and developing countries to attend the General Assembly and Scientific Symposium of URSI.

To qualify for an award the applicant:

1. must be less than 35 years old on September 1 of the year of the URSI General Assembly and Scientific Symposium;
2. should have a paper, of which he or she is the principal author, submitted and accepted for oral or poster presentation at a regular session of the General Assembly and Scientific Symposium.

Applicants should also be interested in promoting contacts between developed and developing countries. Applicants from all over the world are welcome, also from regions that do not (yet) belong to URSI. All successful applicants are expected to participate fully in the scientific activities of the General Assembly and Scientific Symposium. They will receive free registration, and financial support for board and lodging at the General Assembly and Scientific Symposium. A basic accommodation is provided by the assembly organizers permitting the Young Scientists from around the world to collaborate and interact. Young scientists may arrange alternative accommodation, but such arrangements are entirely at their own expense. Limited funds will also be available as a contribution to the travel costs of young scientists from developing countries.

The application needs to be done electronically by going to the same website used for the submission of abstracts/papers. This website is www.papers-GASS2011.ursi.org. The deadline for paper submission for the URSI GASS2011 in Istanbul is 11 February 2011.
A web-based form will appear when applicants check “Young Scientist paper” at the time they submit their paper. All Young Scientists must submit their paper(s) and this application together with a CV and a list of publications in PDF format to the GA submission Web site.

Applications will be assessed by the URSI Young Scientist Committee taking account of the national ranking of the application and the technical evaluation of the abstract by the relevant URSI Commission. Awards will be announced on the URSI Web site in April 2011.

For more information about URSI, the General Assembly and Scientific Symposium and the activities of URSI Commissions, please look at the URSI Web site at: http://www.ursi.org or the GASS 2011 website at http://www.ursigass2011.org/

If you need more information concerning the Young Scientist Program, please contact:

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XXXth URSI General Assembly and Scientific Symposium

Istanbul, Turkey August 13-20, 2011

URSI Student Paper Competition

Chair: Prof. Steven C. Reising
Colorado State University, Fort Collins, CO, USA

Student Paper Prize winners, 1st Place through 5th Place, will be awarded a certificate and check in the amounts of $1500, $1250, $1000, $750 and $500, respectively.

Rules and Guidelines

- First author and presenter must be a full-time university student.
- The topic of the paper must be related to the field of one of the ten URSI Commissions.
- A full paper must be submitted by the abstract deadline. The paper must be not longer than 25 pages and in the single-column, double-spaced manuscript format of the journal *Radio Science*.
- A letter from the student's advisor on university letterhead must be appended to the paper. The letter must state that the author is enrolled as a full-time university student in a degree program. If co-authored, the letter must state that all co-authors played only an advisory role. *No other students are permitted as co-authors*.
- Ten finalists will be chosen based upon quality, originality and scientific merit. They will receive free access to the workshop/short course of their choice. They will be required to attend the banquet, where all finalists will be recognized, and the prizes will be presented.
- The URSI Panel of Judges will consist of the ten URSI Commission Chairs or their authorized representatives, in case of absence.
- In addition, the prizes will be awarded based on the clarity of their presentation, accessibility to the broad audience of the ten URSI Commissions and the ability to answer questions on their work.
- All participants will have the option of submitting their full paper manuscripts for review for publication in a special section of the journal *Radio Science* edited by Prof. Piergiorgio L. E. Uslenghi, Univ. of Illinois at Chicago, IL, USA, 2011 URSI GASS Scientific Program Coordinator.

Characterization of Wireless Channels on the Human Body

Peter S. Hall
Yang Hao

Abstract

This paper gives an overview of the current state of the characteristics of wireless channels on the human body at microwave frequencies. Such channels are increasingly of importance for applications including medical-sensor networks, sports-performance monitoring, and personal entertainment and communications, in which both transceivers are located on the body and exchange data. On-body channels form a subset of body-area-network (BAN) channels. Communications to local base stations off the body and to implants inside the body may also be part of the body-area network, but are not discussed here. The on-body channel is subject to both variation due to the changes of orientation of the antennas as the body moves, to scattering from the body, and to scattering from the local environment. In general, it is difficult to extract the propagation-channel characteristics from those of the wearable antennas. Results are presented for path loss and channel statistics at the 2.45 GHz ISM band and in the 3 to 10 GHz UWB band, and for diversity and MIMO in the ISM band.

1. Introduction

Body-centric wireless communication is now accepted as an important part of the fourth-generation (and beyond) mobile communications systems, taking the form of human-to-human networking incorporating wearable sensors and communications [1, 2]. There are also a number of body-centric-communication systems for specialized occupations, such as paramedics and firefighters, military personnel, and medical sensing and support. Such systems incorporate antennas for communications to base stations off the body, but there are an increasing number of applications where both antennas may reside on the body. In these body-area-network communications channels, propagation will thus involve wave interaction with the body, in addition to scattering from the local environment. A range of operating frequencies are likely to be used in such body-area networks, including low frequencies, such as 10 MHz [3]; the industrial, scientific, and medical (ISM) band, from 2.4 GHz to 2.48 GHz [4, 5, 6]; and the ultra-wideband (UWB) standard, between 3.1 and 10.6 GHz [7, 8, 9]. Much attention has recently been paid to this ISM band due to the increasing number of applications, such as Bluetooth headsets and wristwatch controllers for mobile phones.

This paper gives an overview of studies to date in characterization of channels on the human body. Narrowband characterization – primarily at 2.45 GHz – is described in detail, and reference is made to other frequencies in terms of path-gain dependence in Section 2.1, and higher-order statistics in Section 2.2. Wideband channel characteristics are discussed in Section 3, again in terms of path-loss dependence, and then in terms of time-delay analysis. The properties of multiple antenna systems are covered, including two-antenna diversity at 2.45 GHz and multiple-input multiple-output (MIMO) capacity at the same frequency. Finally, conclusions are drawn, together with a discussion about future needs.

2. Narrowband Channel Characterization

2.1 Path-Gain Distance Dependence

The transmission channel includes both antennas and the propagation channel. Its performance depends on the types of antennas and the orientation of the antennas, as these determine the polarization and the path gain of the various propagation paths that constitute the channel. The distance of the antennas relative to the body also influences the channel, because the gain, radiation pattern, and polarization of the antennas are all altered in proximity to the body. The body

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This is an invited *Review of Radio Science* from Commission B

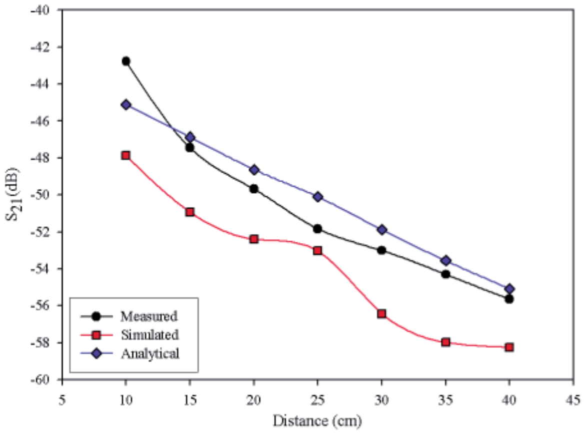

Figure 1. The S_{21} of two monopole antennas mounted on a human body at 2.45 GHz.

size, shape, and posture, and the local environment, all have an effect on what propagation paths are supported and what are their path gains. In addition, movement of the body, both in terms of its relative geometry and its position in the environment, will give rise to variations in the channel.

Radiowave propagation between two antennas on the human body can take the form of a space wave, a surface wave, or a creeping wave. It may experience reflection and diffraction, depending on the positions of the antenna(s), the body geometry, and the environment. It has been shown [10] that the surface wave is predominantly of the Norton type, but is highly dependent on the parameters of the body. This is supported by Figure 1 [11], which shows S_{21} for different distances between two monopole antennas on the chest. This was calculated using the Norton-wave analysis in a simulation using CST finite-element software and a "Norman" full-body numerical phantom, and measurements. Quarter-wavelength monopoles on a 10 cm × 10 cm ground plane parallel to the body's surface were used in the simulation and measurement. Elemental sources were assumed in the analysis. The agreement was good. The surface-wave analysis gives good insight into this propagation mode, but is insufficient for finding the channel performance in other cases. The simulation is numerically demanding, but deals with all propagation modes. Simulation of dynamic channels [12] that model the body's movement by successive simulation of frames from a movie created by an animation software are computationally demanding. Therefore, all of the results given here were derived from measurements on human subjects.

Models for the path-gain (PG) distance dependence at 2.45 GHz have been suggested [4, 5, 6]. Reference [5] described measurements around the body's trunk at chest level for a planar monopole antenna tangential to the body's surface. It discussed several models, including that of [6], and showed that the path gain expressed in dB, P_{dB}, was best given by

$$P_{dB}\left(d\right)=10\log_{10}\left(P_0 e^{-m_0 d}+P_1\right)+\sigma_p n_p \,. \qquad (1)$$

This was based on an exponential decay with the distance, *d*, related to diffraction around a cylinder, followed by a flat saturation point. It included an assumed log-normal variation, since n_p is a zero-mean unit-variance Gaussian random variable. Parameters P_0, P_1, m_0, and σ_p were derived from measured data. The exponential decay rate, m_0, was found to be ≈ 2.0 dB/cm at 2.45 GHz. Reference [4] described the use of quarter-wavelength monopole antennas, described in the above paragraph, and described measurements at 2.45 GHz of channels from the belt to many other positions on the body, including the limbs. Figure 2 shows the path gain as a function of distance for these measurements. Three categories of propagation channels were identified. First, a line-of-sight channel was typified by propagation across the front of the body. Second, there were non-line-of-sight channels, such as from the font of the body to the back, and from the belt to the wrist's outer surface. Finally, there were transitional cases between the first two, such as from the belt to the head. The following

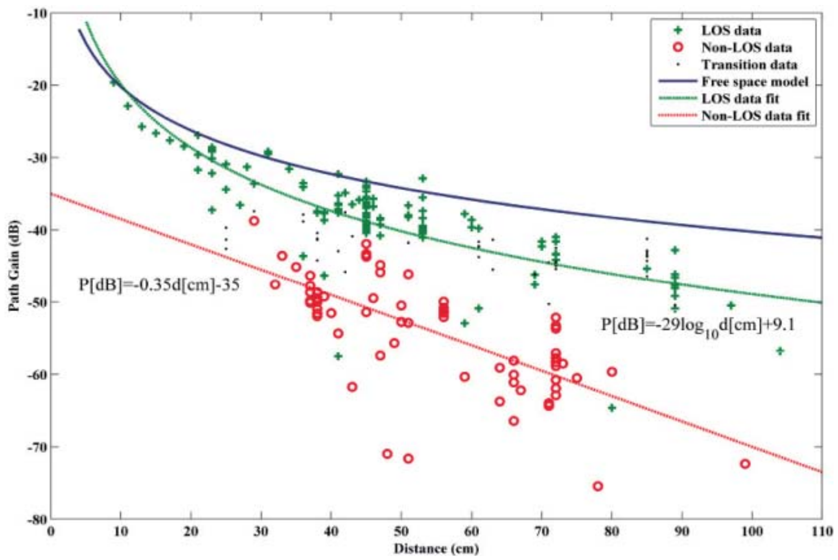

Figure 2. The on-body path gain as a function of distance on the surface of the body. The frequency was 2.45 GHz, and the antennas were monopoles.

expression was obtained for the line-of-sight case:

$$P_{dB}(d) = -29\log_{10} d_{cm} + 9.1 . \quad (2)$$

This shows a falloff of the path gain, P_{dB}, with the distance, d_{cm}, expressed in cm, similar to that in free space but ~ 5 dB lower. The non-line-of-sight cases showed large spread, but could be approximated by an exponential attenuation according to the following linear-regression formula:

$$P_{dB}(d) = -0.35\mathrm{d}_{cm} - 35 , \quad (3)$$

giving a decay rate of 0.36 dB/cm.

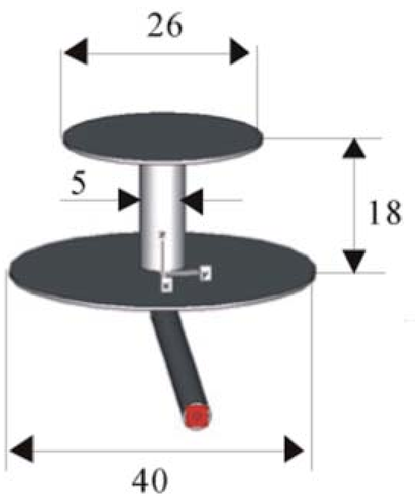

Figure 3a. The top-loaded monopole used for the derivation of statistical models. Metallization on the top of the substrate is shown in black, metallization on the underside is shown in grey, and the substrate boundaries are shown by thin black lines.

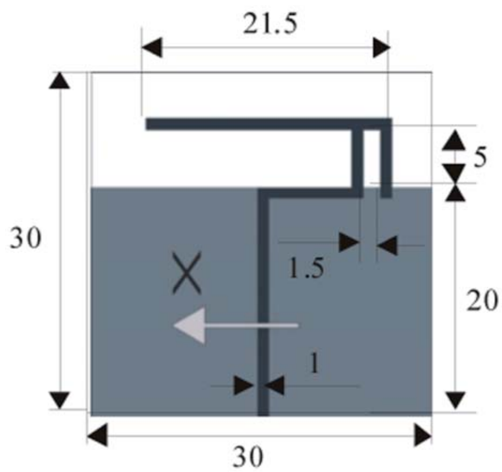

Figure 3b. The PIFA used for the derivation of statistical models (the top-plate height was 5 mm; shading as in Figure 3a).

2.2 Channel Statistics

Statistical characterization of path-gain variations was studied in [13] and [14], where statistical distributions were fitted to the path-gain variation data collected during various activities performed by the subject, such as walking around inside a building, and interacting with doors, other people, and furniture. The channel variations were shown to be non-stationary in [14], which makes derivation of general statistical models by this method difficult. Section 2.2.1 addresses the walking human, and Section 2.2.2 considers the case of naturalistic human movement.

2.2.1 Walking Human

A two-scale approach to modeling random channel variations – which is usually used to characterize mobile-communication channels – was suggested in [15]. Measurements on walking humans were discussed in [16], and are discussed in detail below. It is clear that the

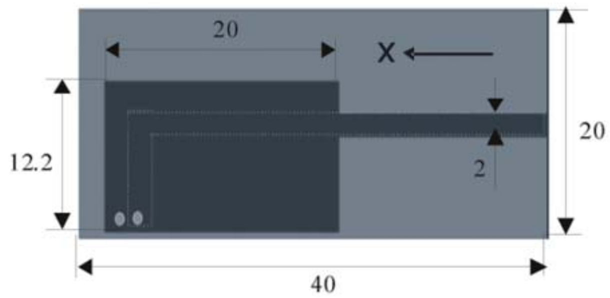

Figure 3c. The printed IFA used for the derivation of statistical models (shading as in Figure 3a).

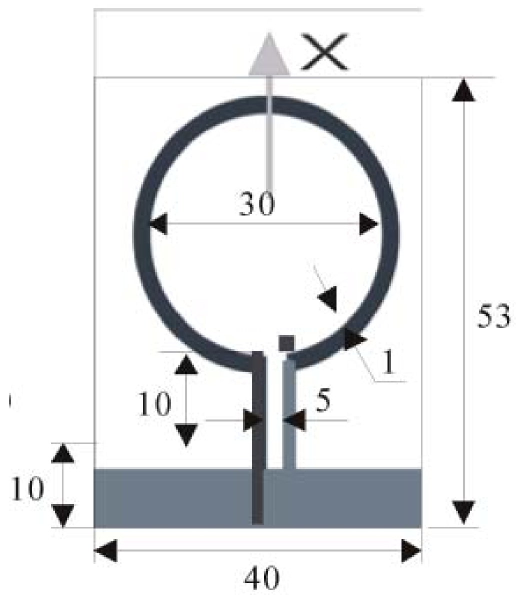

Figure 3d. The printed loop used for the derivation of statistical models (shading as in Figure 3a).

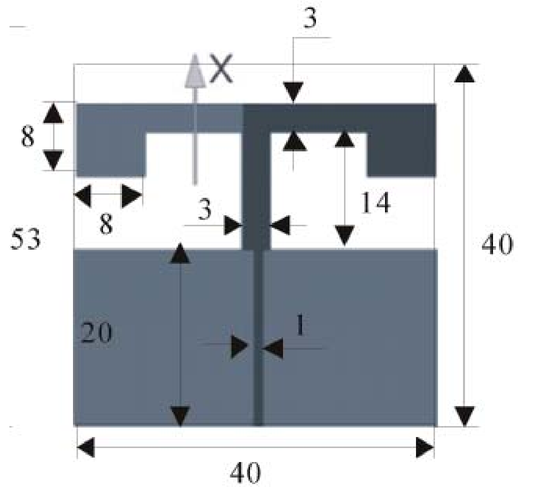

Figure 3e. The printed dipole used for the derivation of statistical models (shading as in Figure 3a).

statistics are dependent on the body, the antenna type, and its orientation. A male and a female test subject were used in various environments and with various antennas operating at 2.45 GHz. The antennas included small top-loaded monopoles, planar inverted-F antennas (PIFAs), printed inverted-F antennas (IFAs), printed loops, and dipoles. The path gain was measured with two identical antennas, for example, PIFA-PIFA or loop-loop. Figure 3 shows these antennas with their dimensions shown in millimeters. All of the printed antennas were made on a 0.8 mm-thick FR4 substrate.

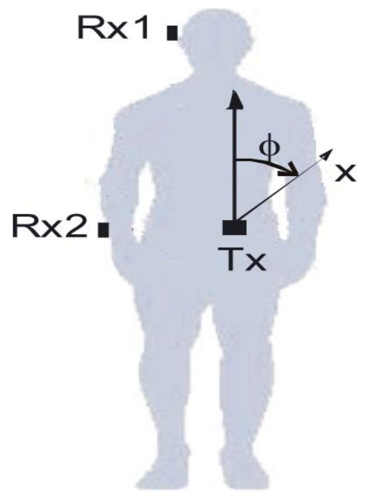

Figure 4. The antenna positions on the body.

Figure 4 shows the antenna positions on the body, indicated by the black boxes. The positions included the left side of the abdomen on the belt, the right side of the head near the ear, and the outer side of the right wrist. In all cases, the antennas were tangential to the skin. Data were gathered for the belt-to-head and the belt-to-wrist channels. In order to examine the effect of antenna orientation on the path gain, data were gathered with some of the antennas in different orientations. The antenna orientation is defined by an angle, φ (Figure 4), between the vertical axis and the antenna's axis, *x*, shown in Figure 3. In the result shown in Table 1, the orientation angle of the transmitting antenna on the belt is referred to as φ', while φ refers to the receiving

Subject	Antennas	φ' (deg)	Short-Term Fading: Rician			Long-Term Fading: Log-Normal	
			s (dB)	σ (dB)	K (dB)	μ (dB)	σ_L (dB)
Male	Monopoles		–3.9	–5.4	–1.5	–56.4	1.9
	Dipoles	0	–2.0	–7.4	2.3	–79.8	1.6
	Loop $\varphi = 140°$	–90	–3.7	–5.5	–1.2	–71.7	2.1
	Loops $\varphi = 50°$	–90	–3.6	–5.6	–1.1	–73.3	2.1
	PIFAs ($\varphi = 140°$)	0	–2.2	–7.3	2.1	–72.9	1.7
		–90	–1.8	–7.8	3.0	–72.3	1.3
		180	–2.1	–7.2	2.0	–72.6	1.3
		90	–3.8	–5.5	–1.3	–75.5	1.9
	PIFAs ($\varphi = -40°$)	0	–2.1	–7.3	2.2	–66.2	1.9
		–90	–1.9	–7.7	2.8	–66.7	1.5
		180	–1.9	–7.6	2.7	–68.6	1.6
		90	–3.8	–5.4	–1.4	–73.7	2.0
	IFAs	0	–3.3	–5.8	–0.5	–67.5	2.2
		–90	–3.5	–5.7	–0.8	–66.9	2.1
		180	–3.6	–5.6	–1.0	–67.9	1.9
		90	–3.7	–5.6	–1.1	–65.6	1.6
Female	Monopoles		–2.8	–6.3	0.5	–56.2	2.1

Table 1. The parameters of the statistical models for short-term and long-term fading in the belt-to-head channel.

antenna. The magnitude of the propagation path gain was measured with a Rhode and Schwartz FSH6 handheld spectrum analyzer with tracking generator, controlled by a small laptop, both carried by the subject in a small backpack. The subject (male or female) performed continuous walking in a heavily cluttered office environment. The spectrum analyzer did a 301-point sweep in 3 s, which corresponded to a 10 ms sampling time. The sweeps were repeated 100 times with a 0.5 s gap between them, thus providing 30100 data points for each measurement. In total, 23 such measurements were performed for two channels for the two subjects, and for a number of antennas and antenna orientations, φ' and φ.

In mobile communications, it is usually assumed that the received-signal envelope can be represented as a product of two terms,

$$r(t) = m(t) M(t), \tag{4}$$

where $m(t)$ is the envelope of the short-term fading (also called small-scale or multipath fading), caused by the interference of several multipath components of the received signal; and $M(t)$ is a term that is subject to long-term fading (also known as large-scale or shadow fading), caused by random shadowing events. The results for continuous walking presented below seemed to confirm that this can also be applied to on-body channels.

The short-term fading term, $m(t)$, was derived by normalizing the measured envelope by the local rms, $M(t)$. Both the short-term and long-term fading data were then re-sampled by taking every third point for short-term fading, and every 22nd point for long-term fading, in order to reduce the correlation between the points to less than 0.5. Seven probability distributions – namely, Rayleigh, Rician, Nakagami, log-normal, Weibull, gamma, and normal – were fitted to the resulting data sets by estimating their parameters with a maximum-likelihood-estimation method. The goodness of fit was then verified by the χ^2 test. Table 1 shows the parameters of the distributions that fit most of the short-term fading and long-term fading data sets derived from the measurements of continuous walking. For short-term fading, the best fit in all cases was Rician, with significance levels greater than 0.1%. The exception to this were the cases of the PIFA and IFA on the wrist, where the significance level for all models was low. In most cases, all the other distributions give significance levels below 0.0001%. Parameters s, σ, and K characterize the strongest ray power, s_2, the average scattered power, $2\sigma^2$, and the ratio of the power in the strongest ray to the average scattered power, $K = s^2 / 2\sigma^2$, respectively. The total mean power, $\Omega = s^2 + s\sigma^2$, was found to be close to 0 dB in all cases, as required due to the rms normalization.

Table 1 also shows the mean, μ, and the standard deviation, σ_L, of the long-term fading envelope, expressed in dB. The range of standard deviations was relatively small, from 1.3 to 2.8 dB. On the other hand, the parameter μ, which represents the average path gain, had large differences. It was highest for the monopoles in both channels, and was at least 10 dB lower for other antennas. The K value for the monopole on the wrist was also high, since the monopoles' polarization normal to the body's surface helped to maintain a strong ray diffracted around the body. PIFAs were the second-best-performing antennas on the wrist, in terms of both the mean path gain and the K value. The polarization of a PIFA is also normal to the body's surface. However, its low profile led to higher shadowing losses and efficiency reductions in the proximity of the body. In the head channel, IFAs gave a mean path gain similar to that of PIFAs when $\varphi' = -40°$, but a large proportion of the received power was due to scattering from the environment. Therefore, IFAs gave a K value about 3 dB lower, and hence, led to stronger fading. The low mean path gain for the IFA, loop, and dipole were believed to be due to their polarizations being parallel to the body's surface. They thus did not couple well to the creeping-wave propagation mode, in addition to their efficiency being significantly reduced by the proximity of the body. The relatively high radiation away from the body in the IFA and loop cases also gave rise to comparable levels of scattering, resulting in K values around 0 dB.

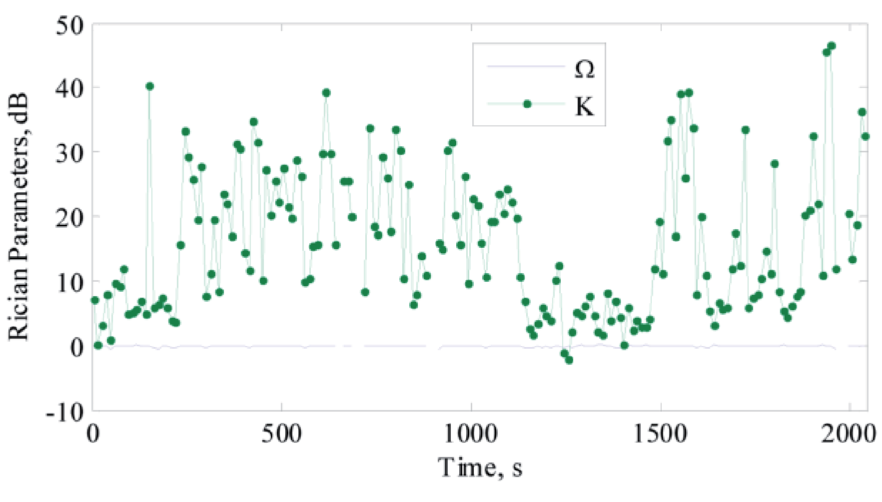

Figure 5. The temporal variation of the Rician parameters Ω and K of the short-term fading for the belt-to-wrist channel in the residential house.

In the second type of measurement, the sweep time of the spectrum analyzer was set to 2 s, and each sweep was taken synchronously with a single pace (two steps) performed by the subject (male). Thus, at the beginning of the time sweep the subject was standing upright, and then started to walk with the left leg and the right arm first moving forward. At the end of the sweep, the subject would return to the standing upright position after moving one pace forward. A hundred repetitions of such paces were undertaken for each of the two channels, both in an anechoic chamber and in the office environment. Two styles of walking were investigated in each case: walking with natural arm movement, and walking with the arms held against the sides of the body. Only monopole antennas were used in these measurements. There were thus a total of eight such measurements, which we will refer to as synchronous-walk measurements.

The power spectral density was extracted from repetitive single-pace walks. For the belt-to-wrist channel, the male showed higher spectral-content power, due to the female walking with very little arm swinging. A component close to 1 Hz and its first harmonic near 2 Hz were present for both the male and female, and corresponded to the arm-swinging rates. Unlike the other antennas, the monopoles produced a spectrum with the second harmonic stronger than the first. This can be explained by the radiation pattern of the monopoles, which led to the signal significantly decreasing not only when the arm with the antenna was behind the body, as for all the other antennas, but also when the arm was in front. These spectral components were much less prominent in the head channel. It can therefore be concluded that the spectrum depends on the channel and the subject's movements, and for the links with limbs, it will contain peaks at the rate of walking.

The autocorrelation of the magnitude data in each sweep was calculated. The correlation coefficients for all the head-channel measurements were found to be low and very similar. In the wrist channel, there were strong periodic features that were dependent on antenna type. For a correlation of less than 0.5, the correlation length was less than 20 ms for the head channel, and up to 30 ms for the wrist. For long-term fading, there were high peaks, up to values of 0.3 at a 1 s time lag, which were due to the walking action. For a correlation of less than 0.5, the correlation length was less than 220 ms.

Environment	μ_{kdB}	σ_{kdB}
Office	9.4	8.1
House	15.7	10.9
Outdoor	8.2	3.7
Car	14.5	7.1

Table 2. The statistical parameters representing the variation of Rician short-term fading parameters (belt-to-wrist channel, 2.45 GHz).

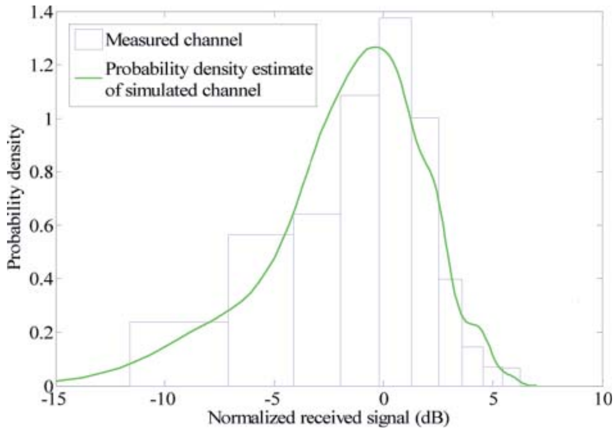

Figure 6a. A comparison of the normalized measured channel with the simulated auto-regression model output for the left-waist-to-right-head on-body channel whilst the user was mobile in an anechoic chamber.

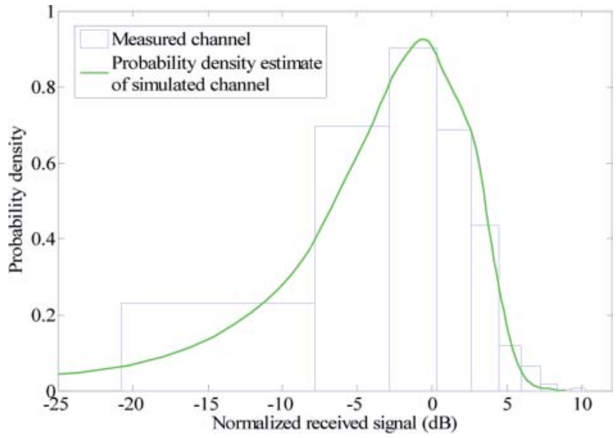

Figure 6b. A comparison of the normalized measured channel with the simulated auto-regression model output for the left-waist-to-right-head on-body channel whilst the user was mobile in an open office area.

2.2.2 Naturalistic Movement

In naturalistic human movement, the statistics described in the previous section become non-stationary. This posed some problems in deriving simple-to-use models. In [17], the two-scale model was extended to the case of natural activities in a number of different environments. Although it was found that no theoretical formula fitted the data well, approximate probability distributions were calculated. Figure 5 shows the temporal variation of the Rician parameters for the belt-to-wrist channel of the subject while performing a variety of activities in a residential house, such as cooking, eating, and watching TV. The Rician PDF is given by the formula

$$p_R(r) = \frac{2r(K+1)}{\Omega}\exp\left[-\frac{(K+1)r^2}{\Omega}-K\right]I_0\left[r\sqrt{\frac{K(K+1)}{\Omega}}\right] \quad (5)$$

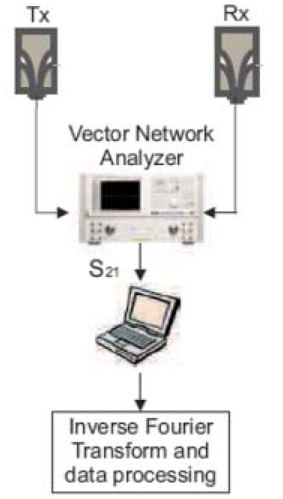

Figure 7a. The setup of the on-body UWB measurement.

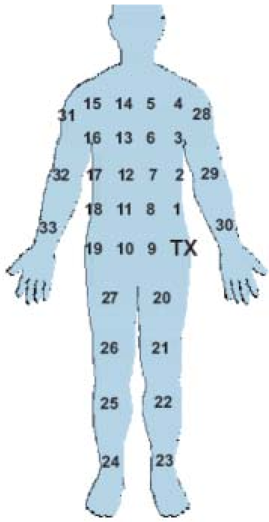

Figure 7b. The location of the transmitting antenna and the 33 positions of the receiver.

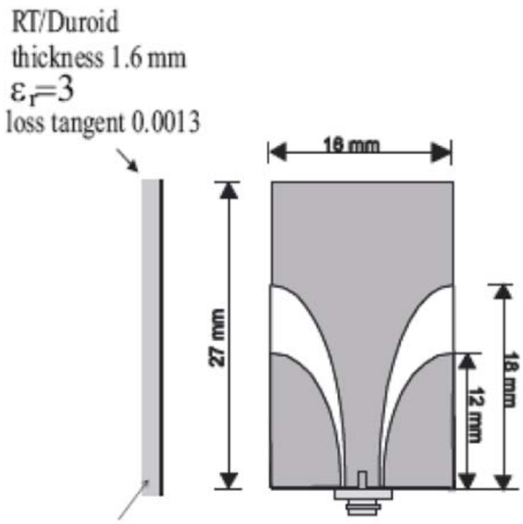

Figure 7d. The layout of the tapered-slot antenna.

where Ω is the mean power, and *K* is the ratio of the power of the strongest propagation path to the average power of the random multipath component. It could be seen that the *K* parameter, expressed in dB ($K_{dB} = 10\log_{10} K$), was greater than 0 dB most of the time. The parameters μ_k and σ_k of the log-normal distributions, fitted to the Rician *K* values, are summarized in Table 2 for the various measurement scenarios and the belt-to-wrist channel. The parameters were expressed in dB, so that $\mu_{kdB} = (20/\ln 10)\mu_k$, and $\sigma_{kdB} = (20/\ln 10)\sigma_k$. The table shows *K* values that were quite large in most cases, and therefore a strong dominant path must usually have been present. There was marked non-stationarity in the indoor environments, and much less in the outdoor or in-car environments, which was attributed to the fact that the activities in these environments were rather limited and lacked diversity.

In [18], auto-regression (AR) transfer coefficients were obtained from the autocorrelation of measured time series data at 2.45 GHz for a number of on-body channels. The coefficients enabled creation of new time-series data that well represented the actual channels. The statistics of these simulated channels, using 10 coefficients, were compared with measurements. Figure 6 shows the comparison of the empirical histogram of the entire measured channel normalized to its global mean, and the estimated probability density function of the auto-regression-based simulation, for two environments. For the anechoic environment, the parameters μ and σ of the measured (subscript *m*) and simulated (subscript *s*) results were $\mu_m = 2.39$, $\sigma_m = 1.11$ and $u_s = 2.57$, $\sigma_s = 1.10$, respectively. For the open office environment, they were $\mu_m = 1.35$, $\sigma_m = 1.20$, and $u_s = 1.57$, $\sigma_s = 1.16$, respectively. The good agreement confirmed the validity of the approach.

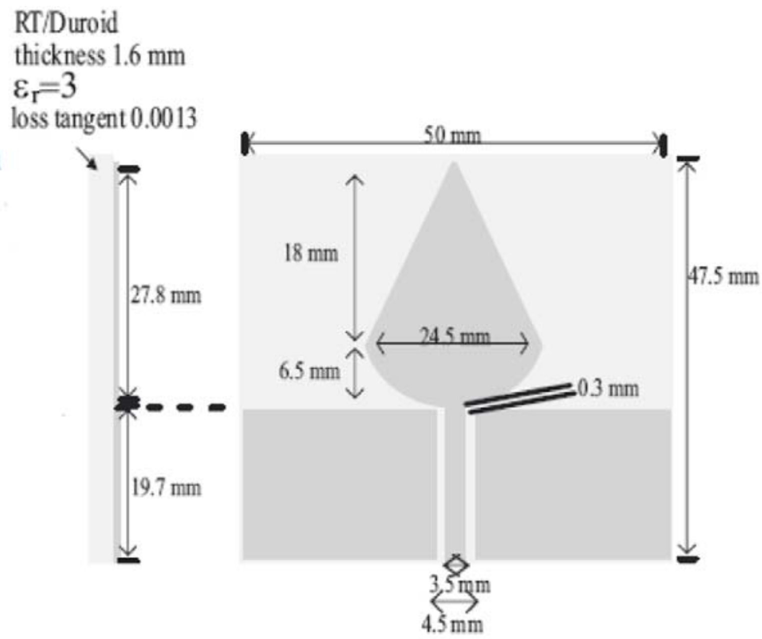

Figure 7c. The layout of the planar inverted-cone antenna.

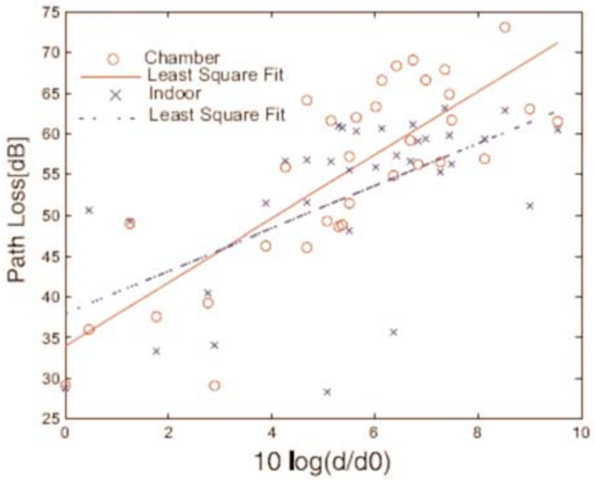

Figure 8a. The measured and modelled UWB path loss for the on-body channel as a function of the logarithmic transmitter-receiver separation distance for modified coplanar waveguide-fed planar inverted-cone antennas (PICA).

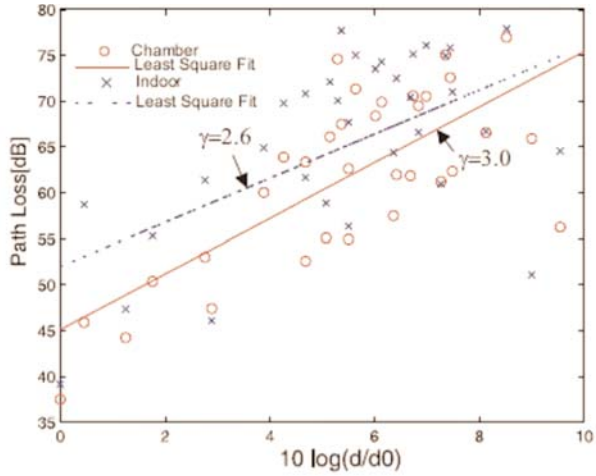

Figure 8b. The measured and modelled UWB path loss for the on-body channel as a function of the logarithmic transmitter-receiver separation distance for tapered-slot antennas (TSA).

3. Wideband Channel Characterization

3.1 Introduction

Wideband channel characterization has been centered on the ultra-wideband (UWB) standard, covering the 3 to 10 GHz band. Many efforts have been made to characterize the UWB on-body radio channel [19-23]. In [19], a measurement campaign was performed using two different antennas exhibiting different radiation characteristics. In [21], the propagation around the trunk was fully analyzed, and a statistical model was proposed. In [23], the effect of the indoor environment was studied without considering the impact of antenna-radiation characteristics. To illustrate the issues in such work, results from [24] are now described.

A measurement campaign in the UWB frequency band, 3-10 GHz, was performed using a pair of miniaturized coplanar-waveguide- (CPW) fed tapered-slot antennas (TSA) [25], as well as a pair of modified coplanar waveguide-fed planar inverted-cone antennas (PICAs) [26], as shown in Figure 7. Free-space and body-mounted characterizations of the aforementioned antennas were presented in [27]. To enable modeling and prediction of the time-delay characteristics of the radio channel, the measured data were fitted to empirical statistical models. The goodness of different statistical distributions in fitting the root-mean-square delay spread was evaluated. The results demonstrated that the Nakagami model provided the best fit. When measurements were carried out in the indoor environment, more reflected components were collected, causing a degradation of the goodness of fit of the statistical model. Due to its more directive radiation [27], the miniaturized antenna (TSA, tapered-slot antenna) was

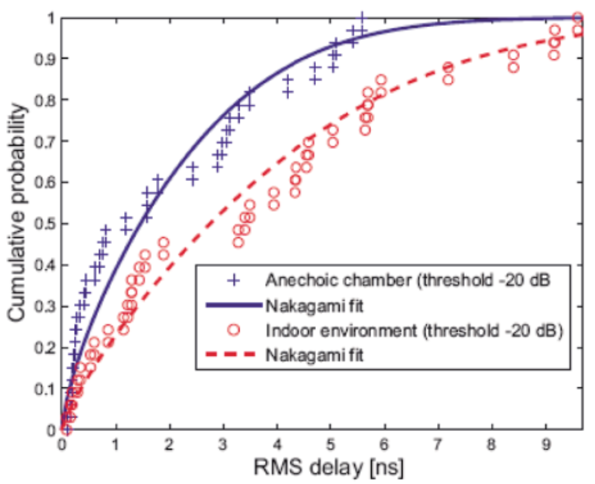

Figure 9a. The cumulative distribution of the rms spread delay fitted to a Nakagami distribution applying a threshold of -20 *dB with a coplanar waveguide-fed planar inverted-cone antenna (PICA).*

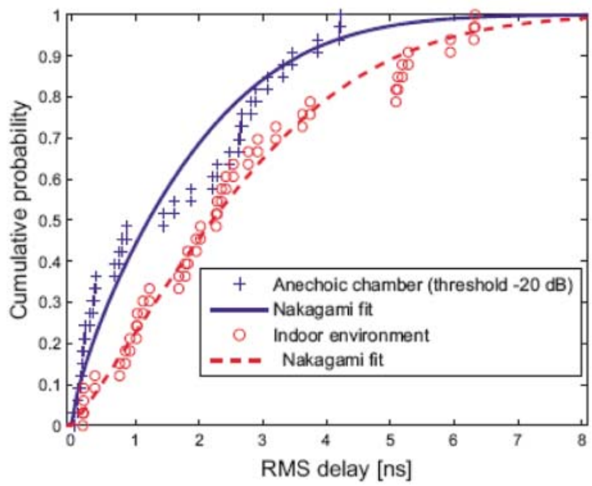

Figure 9b. The cumulative distribution of the rms spread delay fitted to a Nakagami distribution applying a threshold of -20 *dB with a coplanar waveguide-fed tapered slot antenna (TSA).*

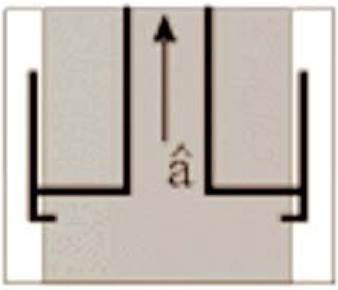

Figure 10a. Configuration A of the printed-IFA diversity antenna. The black lines are the metallic conductors. The metallic ground planes are in grey, and the substrate size is indicated by the thin line.

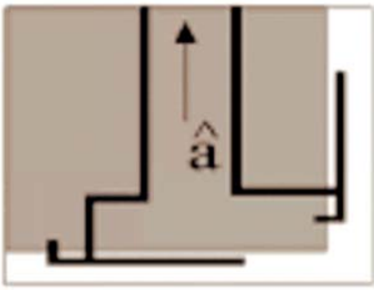

Figure 10b. Configuration B of the printed-IFA diversity antenna. The black lines are the metallic conductors. The metallic ground planes are in grey, and the substrate size is indicated by the thin line.

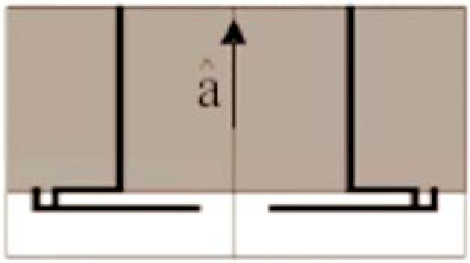

Figure 10c. Configuration C of the printed-IFA diversity antenna. The black lines are the metallic conductors. The metallic ground planes are in grey, and the substrate size is indicated by the thin line.

less affected by the reflections from the human body and the indoor environment. It showed improved time-delay behavior, and better fit the statistical model.

Figure 7a shows the measurement setup. Figure 7b shows the location of the two antennas during the measurements: the transmitting antenna was placed on the left side of the belt, while the receiver was moved along the front part of the body. Thirty-three different positions were measured (on the chest, legs, and arms) to ensure sufficient data collection for channel characterization and modeling.

3.2 Path Loss

The path loss was directly calculated from the measured data, averaged over each frequency point [28], and it was represented as a function of the distance between the transmitter and receiver, as done in [29]. The path-loss exponent was used to describe how fast the received power decayed with the distance. The exponent was found to be 3.9 for the planar inverted-cone antenna, and 3.0 for the tapered-slot antenna (Figure 8). When measurements were performed in the indoor environment, the reflection from the surrounding scatterers increased the received power, causing reduction of the path-loss exponent to 2.6 for the planar-inverted-cone-antenna case, and to 2.6 for the tapered-slot antenna. The planar inverted-cone antenna, having a more-omnidirectional radiation pattern than the tapered-slot antenna, was more affected by multipath reflections, and the reduction in the exponent was more significant. The exponent values agreed with the values presented in [22], which were 3.3 in free space and 2.7 in the office environment. To improve the accuracy of the path-loss model, a zero-mean, normally distributed statistical variable was introduced to represent the deviation of the measurements from the calculated average path loss. In the anechoic chamber, the standard deviation was $\sigma = 6.8$ for the planar inverted-cone antenna, and $\sigma = 8.2$ for the tapered-slot antenna.

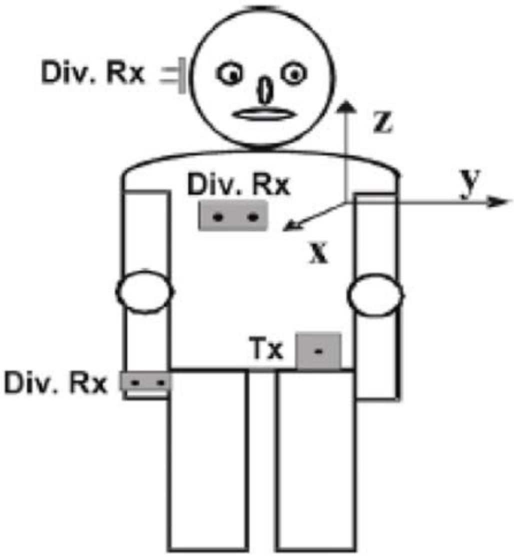

Figure 11. The body-fixed reference system.

Orientation		MRC (dB)	ρ_{env}	Mean Power (dB)	Power Difference (dB)
Config A	Or1 ($\hat{a} = \hat{z}$)	8.3	0.59	–61.08	0.28
	Or2 ($\hat{a} = \hat{x}$)	8.6	0.32	–69.50	0.92
	Or3 ($\hat{a} = -\hat{z}$)	7.2	0.36	–68.21	3.54
	Or4 ($\hat{a} = -\hat{x}$)	6.8	0.85	–50.81	0.80
Config B	Or1 ($\hat{a} = \hat{z}$)	11.1	0.18	–66.39	0.76
	Or2 ($\hat{a} = \hat{x}$)	4.5	0.62	–36.51	2.43
	Or3 ($\hat{a} = -\hat{z}$)	6.5	0.79	–40.43	1.54
	Or4 ($\hat{a} = -\hat{x}$)	9.7	0.27	–65.25	3.44
Config C	Or1 ($\hat{a} = \hat{z}$)	7.9	0.41	–58.32	2.82
	Or2 ($\hat{a} = \hat{x}$	7.9	0.45	–35.97	5.25
	Or3 ($\hat{a} = -\hat{z}$)	6.0	0.65	–39.11	1.70
	Or4 ($\hat{a} = -\hat{x}$)	7.7	0.43	–58.11	2.36

Table 3. The performance of diversity antennas in the belt-to-head channel (the frequency was 2.45 GHz, and the configuration and orientation were as shown in Figures 10 and 12, respectively).

Figure 12. The orientation of the printed IFA on the body for a front view of the body: the four antennas on the left are the receiving diversity antennas, and the antenna on the right is the transmitting antenna on the belt, with $\hat{a} = -\hat{y}$ for the belt-chest, belt-head, and belt-back channels, $\hat{a} = -\hat{z}$ for the belt-wrist channel, and $\hat{a} = \hat{y}$ for the belt-ankle channel.

3.3 Time-Delay Analysis

The time-domain dispersion of the received signal strongly affects the capacity of UWB systems [30]. This effect was characterized by the first central moment (mean excess delay, τ_m), and the square root of the second-order central moment (root mean square, τ_{rms}) of the power-delay profile (PDP) [31]. Using the Akaike information criteria, it was found that the Nakagami distribution provided the best fit for the rms delay-spread data. Figure 9 shows the cumulative distribution of the delay spread obtained applying a threshold of -20 dB fitted to a Nakagami distribution.

The tapered-slot antenna, due to its directional radiation characteristics, was less affected by reflections from the human body and surrounding scatterers. When measurements were performed in an indoor environment, the multipath effect produced higher delay spread. The planar inverted-cone antenna picked up more multipath components due to its omnidirectional radiation pattern.

4. Multiple Antenna Systems

Multiple antenna systems are used widely in personal communications systems to provide fade mitigation and increased capacity [32]. These are generically known as multiple-input multiple-output (MIMO) antenna systems. MIMO utilizes diversity to enhance system performance through increasing the signal power or transmission rate. Space or pattern diversity, using a pair of antennas at a base station, is widely used to increase the average received power at the handset, and thus overcome the potentially disastrous effect of deep fades due to multipath fading. Arrays of antennas at either end of a link can make use of the multipath environment to enhance the data rate using time-space coding methods. The potential of such methods has been examined for on-body channels, and results are described in this section.

4.1 Diversity

Good diversity performance requires the signals received at the multiple-diversity antennas to be relatively uncorrelated and approximately equal in magnitude. For on-body communication channels, pattern- and polarization-diversity antennas must be carefully designed to prevent one branch antenna signal being dominant compared to the other for various positions and postures of the body. Otherwise, the high power imbalance can severely affect the diversity performance. Due to limited changes in the antenna orientations for most of the on-body channels, the power imbalance will be larger if the pattern or polarization of the diversity antennas is different. Space diversity is thus considered as a good choice for most on-body channels. In general terms, the diversity gain is dependent on the types and relative orientations of the antennas, the channel being considered, and the movement of the body. The improvement offered by diversity for the on-body channels was reported in [33-36] using monopole antennas. A comparison of the space and pattern diversity for on-body channels was given in [36]. Antenna details and results at 2.45 GHz are given below, whilst those for 5.8 and 10 GHz were given in [37].

As shown in Figure 3, IFAs were combined in space-diversity configurations (A and C in Figure 10) on substrate sizes of 40 mm × 40 mm and 40 mm × 60 mm and with ground-plane sizes of 30 mm × 40 mm and 30 mm × 60 mm,

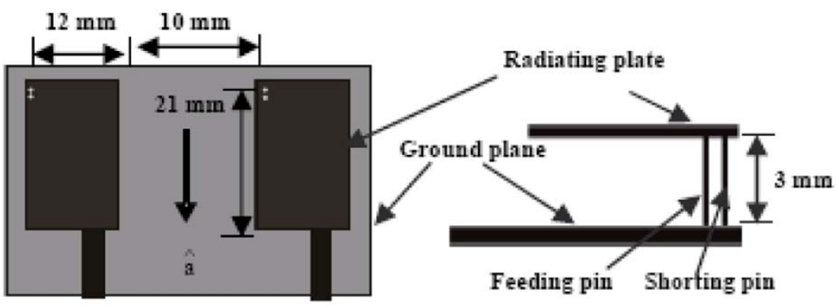

Figure 13. Top and side views of the PIFA array.

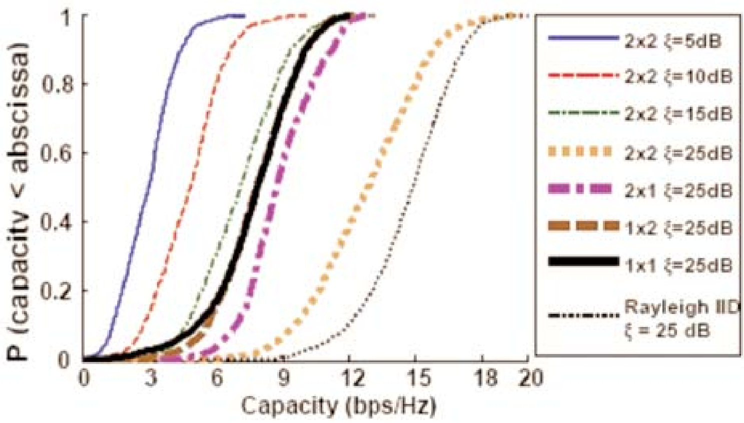

Figure 14. The capacity cumulative distribution functions for the belt-head channel.

respectively. Configuration B was a space-, pattern-, and polarization-diversity antenna, and of the same substrate size as A and with a ground-plane size of 35 mm × 35 mm. The mutual coupling between the IFAs was 16, 12, and 10 dB for configurations A, B, and C, respectively. The transmitting antenna is shown in Figure 12e. The unit vector, $\hat{a}$, pointed towards the direction of the feeding line is shown in Figure 10, and $\hat{x}$, $\hat{y}$, $\hat{z}$ are the unit vectors as shown in Figure 11. For each channel, the transmitting antenna was oriented such that its main-beam direction faced the receiving antennas. The three diversity configurations, shown in Figure 10, were oriented in different ways, as shown in Figure 12.

Measurements were taken in an indoor environment in a typical laboratory, which was an L-shaped room containing equipment, tables, chairs, and computers, providing a rich multipath propagation environment. The size of one arm was 5 m × 2 m, and the other arm was 7.5 m × 1.5 m. The two receiving antennas were connected to the two ports of a calibrated vector network analyzer, operating in tuned dual-channel receiver mode. During the measurements, random activities, such as walking, moving hands, eating, bending down, etc., were performed, and the same activities were repeated for every measurement.

Some sample results for the belt-to-head channel are shown in Table 3, where the maximal ratio-combining diversity gain, envelope correlation, mean channel received power, and branch power difference are given. In general, useful diversity gains were obtained. Configuration B, which gave space, pattern, and polarization diversity, gave the best gain, but only in orientation 1. In this case, the signal correlation was low, as was the channel mean power and the received branch power difference. On the other hand, orientation 2 for the same antenna configuration gave high mean power, and larger correlation and power differences. Results were also obtained for monopole and planar inverted-F antennas, as shown in Figure 3. In general, these showed higher mean powers due to their polarization normal to the body's surface, and good diversity gains that were less dependent on configuration and orientation. However, they both had thicker profiles than the IFA, and thus would result in larger terminals.

4.2 Multiple-Input Multiple-Output Antenna Systems

Multiple-input multiple-output (MIMO) systems have been studied extensively for mobile and personal communication links [39-41]. The potential capacity increase for on-body channels offered by MIMO systems was examined in [42], and the methodology and results are summarized in this section. Due to a general perception of relatively small levels of multipath on the human body, MIMO systems have been expected to not be useful. However, the results below showed useful increases, due to the combination of body and environmental scattering at 2.45 GHz. In addition, the capacity increase in Rician fading channels depends on the degree of decorrelation offered by the scattering environment providing the multipath richness, and also on the SNR (signal-to-noise ratio). Body movement may also change the fading distribution of the spatially separated sub-channels when multiple antennas are used. For a fixed transmitted power, the presence of line-of-sight means a high signal-to-noise ratio (SNR) at the receivers, which may increase the channel capacity compared to a non-line-of-sight (NLOS) link with the same configuration. However, on the other hand, line-of-sight links have a low degree of scattering, which introduces a high correlation among the spatial sub-channels. For a specific value of SNR, i.e., variable transmitted power, line-of-sight can thus decrease the channel capacity compared to a non-line-of-sight link at the same level of SNR. Hence, there is a tradeoff between the effect of increased SNR or increased correlation on the channel capacity. It was shown in [43] that at high SNR values, the reduction in capacity due to an increase in correlation is overcome by the high SNR.

To characterize 2 × 2 narrowband MIMO channels for on-body wireless channels, measurements were performed in a way similar to those in Section 4.1, using an array of two microstrip-fed planar inverted-F antennas (PIFA) on an 0.8 mm thick FR4 substrate (as shown in Figure 13) at both the transmitter and receiver locations. The ground plane size was the same as the substrate size, which was 45 mm × 40 mm. The thickness of the radiating plate of the PIFA was 1 mm, and the distance between the short-circuit pin

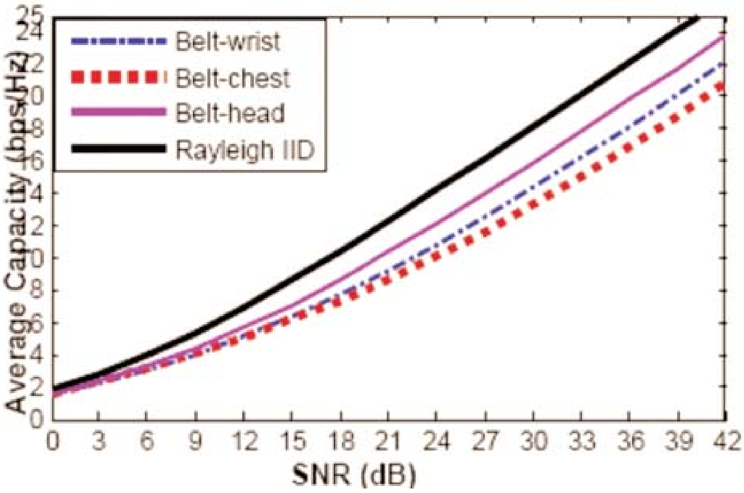

Figure 15. The average capacity as a function of the SNR.

and the feeding pin was 3 mm. The other dimensions of the antenna are shown in Figure 2. The reflection coefficient of the PIFA antenna elements was below -12 dB. The mutual coupling between the two PIFA elements was -12.5 dB. Measurements were performed in an indoor environment, which was a 7.5 m × 9 m laboratory, containing equipment, tables, chairs, and computers, thus providing a rich multipath propagation environment. The two transmitting antennas were connected through an RF switch to a signal generator operating at 2.45 GHz. The switching time of the RF switch was 40 µs, which was much less than the coherence time of the channel [37]. The two receiving antennas were connected to the two ports of a vector network analyzer (VNA), calibrated in tuned receiving mode, with a single-frequency sweep at 2.45 GHz.

The presence of a strong ray introduces strong correlation among the sub-channels [41]. The belt-chest channel was a good example of this, for which the direct ray was much stronger than the multipath components, and hence the sub-channels were highly correlated. A high correlation among the MIMO sub-channels reduced the throughput gain, and thus less improvement in the channel capacity was observed. By way of illustration, the spatial correlation matrix of the belt-head channel is given below:

$$\rho = \begin{bmatrix} 1 & \rho_{11}^{12} & \rho_{11}^{21} & \rho_{11}^{22} \\ \rho_{12}^{11} & 1 & \rho_{12}^{21} & \rho_{12}^{22} \\ \rho_{21}^{11} & \rho_{21}^{12} & 1 & \rho_{21}^{22} \\ \rho_{22}^{11} & \rho_{22}^{12} & \rho_{22}^{21} & 1 \end{bmatrix} =$$

$$\begin{bmatrix} 1 & 0.67+j0.39 & 0.31-j0.27 & 0.27-j0.02 \\ 0.67-j0.39 & 1 & 0.13-j0.36 & 0.17-j0.24 \\ 0.31+j0.27 & 0.13+j0.36 & 1 & 0.56+j0.33 \\ 0.27+j0.02 & 0.17+j0.24 & 0.56-j0.33 & 1 \end{bmatrix} \quad (6)$$

The correlation between the two transmitting signals, ρ_{11}^{12} and ρ_{21}^{22}, was high, and this was so for all channels. This might have been due to the absence of local scatterers in the near vicinity of the transmitting antennas, as the transmitting array was mounted on the waist position and there was less movement of the hands and other body parts near it to cause any significant local scattering. The correlations between the received signals at the two receiving antennas, ρ_{11}^{21} and ρ_{22}^{12}, were low for all the cases except for the belt-chest channel, which was a more-static channel compared to the other two. The line-of-sight component was much stronger, and was only occasionally shadowed by the movement of the hands. The positioning of the receiving antennas on the body for other two channels was such that the antennas were surrounded by a large number of local scatterers in the form of moving body parts. These channels also involved rapid movement of the receiving antennas in the environment, as well as with respect to the transmitter: thus, a high degree of scattering resulted in lower correlation. This low correlation showed that the body movement and the relative movement of the transmitting and receiving antennas significantly decorrelated the sub-channels. The same sort of result is reported in Section 4.1, where the correlation between the received diversity-branch signals was quite low, resulting in high diversity gains. The correlation coefficients, ρ_{22}^{11} and ρ_{21}^{12}, between the sub-channels that did not share any antenna were very low in all cases, except for the belt-chest channel, in which case the correlation coefficient was moderate. In general, correlation among the sub-channels for the three on-body channels was not significantly high, apart from the belt-chest channel.

The cumulative distribution function (CDF) of the channel capacity as a function of the average received SNR, ξ, is shown by thin lines in Figure 14 for the belt-head channel. Also shown are the cumulative distribution functions for other configurations, i.e., MISO, SIMO, and SISO at $\xi = 25$ dB, which are represented by thick lines. The capacities were similar for the belt-wrist and belt-chest channels. Despite the strong correlation among the sub-channels and the presence of a strong direct link due to line-of-sight for the belt-chest channel, the improvement in capacity offered by MIMO over the same channel with SISO, MISO, and SIMO links was noticeable. The capacity improvement might have been due to a number of factors. The most dominant was the multipath richness of the environment. Although there was strong line-of-sight, the direct ray due to the creeping wave was attenuated while propagating on the surface of the body. This fact and the presence of a rich scattering environment meant that the Rician K factor was not as high as expected. The other reason might have been due to the fact that the communication was short range, and the assumption of plane wavefronts may not have been valid. This would mean that the spherical wavefront was being exploited to achieve high capacities [40].

The variation of average capacity with SNR (in dB) is shown in Figure 15 for the three channels. For all the three channels, the increase of the average capacity with each 3 dB increase in SNR was less than 2 bps/Hz. This should be compared to [39], where it was shown that in the high-SNR regime, the increase in capacity with every 3 dB increase in SNR for an independent $n \times m$ MIMO system was $\min(n, m)$ bps/Hz. However, in general, the values were not too far away from 2 bps/Hz for the belt-head and belt-wrist channels, where the correlation was comparatively low.

5. Conclusions

The characterization of channels on the human body is becoming more significant as more applications involving body-worn sensors and networks appear. On-body channels are very much dependent on the types of antennas and on body motion. The very difficult simulation challenges mean that measurements are necessary to the characterization process. This paper has described a number of studies in deriving channel properties and models. These were made primarily at 2.45 GHz, where many of the current body-area networks are operated, and in the ultra-wideband 3-to-10 GHz band, which shows much promise for these applications. In both bands, multipath existed on the body but was very much posture dependent. In general, the channel was subject to fast fading due to body and environmental scattering. In addition, shadowing by the body was present, but its significance could be offset by environmental scattering. In the narrowband case, fades of more than 50 dB could be seen. Models of path loss as a function of distance were obtained. Statistical models for very regular motion, such as walking, were often Rician for fast fading, and log-normal to describe shadowing. In natural human movements, the statistics were non-stationary. Where high levels of multipath occurred, diversity gain was useful for most channels, and MIMO techniques yielded good capacity increases. Further studies to examine the effect of different body sizes and shapes are needed, and also to consolidate characterization for a wider range of body activities and frequency bands. Although simulation was not discussed in this paper, the dynamics of the human body pose enormous computational challenges that need to be addressed.

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Overview of Performance Lower Bounds for Blind Frequency-Offset Estimation

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Abstract

This paper focuses on performance bounds for estimating the frequency and the phase of a received signal when the complex amplitude of the signal is non-constant and unknown. Receivers need to perform such an estimation in many application fields, including digital communications, direction-of-arrival estimation, and Doppler radar. While in digital communications the non-constant complex-signal amplitude is a discrete random variable related to the transmitted information bits, in many other signal-processing fields this non-constant amplitude is typically modeled as multiplicative Gaussian noise. Fundamental lower bounds on the mean square error of any frequency-offset and phase-shift estimator are continuously employed in all these application fields. They serve as a useful benchmark for judging the performance of practical estimators. We present an overview of such bounds with their respective areas of interest, and their associated derivations in closed form.

1. Introduction and Motivation

Let us consider digital bandpass communication over an additive white Gaussian noise (AWGN) channel, using linear modulation. An information bit sequence is first channel-encoded, and then mapped to a block of complex numbers (data symbols) belonging to a discrete symbol constellation set, Ω. The *channel encoder* introduces structured redundancy in the transmitted bit sequences: this makes it possible to detect and correct at the receiver some of the bit errors that have occurred. *Symbol mapping* is performed to improve the bandwidth efficiency. The resulting data symbols are first applied to a square-root Nyquist transmitting filter, and then multiplied to a sinusoidal transmitting carrier signal in order to obtain a signal that is suitable for transmission over the bandpass channel. At the receiver end, the received signal is multiplied with a carrier signal matched to the transmitted carrier signal, applied to a filter matched to the transmitting filter, and sampled at the correct instants in time.

To enable reliable detection of the transmitted information bits from the resulting observation samples, it is imperative that the carrier signals at the transmitter and the receiver have almost exactly the same frequency and phase. However, as the carrier oscillators at the transmitter and receiver operate independently, their frequency and phase are not the same. The demodulation at the receiver is performed using a local reference carrier signal that exhibits a frequency offset, φ_1, and a phase shift, $2\pi\varphi_0$, vis-à-vis the received modulated carrier signal. In this case, the observation samples can be modeled as a noisy version of a complex sinusoid with frequency φ_1 and phase $2\pi\varphi_0$ and with a non-constant complex-valued amplitude equal to the unknown transmitted data symbols or realizations of a multiplicative Gaussian noise process. In order to cope with the unknown parameters, φ_0 and φ_1, the receiver is fitted with an estimation unit, which has to estimate the quantities φ_0 and φ_1 from the observation samples. Once the frequency offset and the phase shift have been estimated, the demodulated signal is *corrected* in order to compensate for them. The detection unit of the receiver subsequently decides upon the received information bits based on the corrected observation samples, assuming perfect frequency-offset and phase-shift compensation. A result of the latter assumption is that the accuracy of the estimation unit has direct repercussions on the accuracy of the detection unit.

Aside from a mismatch between the transmitting and receiving carrier frequency, the frequency offset, φ_1, can also result from the Doppler effect. If a vehicle is transmitting information to the receiver side and is simultaneously moving, the transmitted carrier frequency is modified by

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the Doppler effect, and the receiver is not well adapted in frequency. This Doppler effect – which is a drawback in digital communications – can be of great interest in some applications. For instance, radar based on the Doppler effect is able to find the velocity of a target. In other applications, such as direction-of-arrival (DOA) estimation, the spatial frequency related to the angle-of-arrival in an array processing can be mathematically seen as a carrier-frequency offset. As a consequence, besides digital communications, there are a lot of applications for which estimating a frequency disturbed by a non-constant amplitude is needed. Unlike digital communications, this non-constant amplitude is not associated with information bits, but with other parameters, such as the Doppler spread for Doppler radar, or the spatial distribution of the source for direction-of-arrival estimation [1-3].

The estimation accuracy is usually measured by the mean square estimation error (MSEE). This is the expected value of the squared difference between the estimated and the true values of the frequency offset and the phase shift. The estimation unit that minimizes the mean square estimation error is referred to as the *minimum mean square error* (MMSE) estimator. In many practical situations, minimum mean square error estimation gives rise to a prohibitive computational burden, and one has to resort to approximation techniques. The various existing estimation units are the result of applying these techniques (see, e.g., [4]).

Rapid developments in digital communications [5-8] and signal-processing applications [1-3] have caused a nonstop increase in the requirements that are imposed on the estimation units' design. This has also provided a constant impulse to the research on the fundamental lower bounds on the attainable estimation accuracy (see, e.g., [1-3, 9-20, 20-31]). On the one hand, such bounds serve as a useful benchmark for judging the performance of practical estimators. On the other hand, if interpretable closed-form expressions exist, they also might provide useful insight into the influence of the various signal parameters on the achievable estimation accuracy.

In this tutorial, we focus on the derivation and the analysis of such bounds. One of the most celebrated performance limits is the Cramer-Rao bound (CRB) [32], which is known to be a tight bound for a wide class of estimators, provided that the SNR (signal-to-noise ratio) is sufficiently high. In the applications considered, the statistics of the observation samples depend not only on the frequency offset and the phase shift to be estimated, but also on the statistics of the non-constant amplitude. This makes the computation of the Cramer-Rao bound far from trivial. In order to avoid the computational complexity associated with the true Cramer-Rao bound, several alternative Cramer-Rao-like bounds have also been proposed in the literature (see, e.g., [2, 9, 10, 14, 26, 31]). We present an overview of these bounds with their respective areas of interest and their associated derivations in closed-form for various cases (coded/non-coded digital modulation, circular/non-circular multiplicative noise). It is well known that the Cramer-Rao bound (and, in particular, the Cramer-Rao bound for frequency-offset estimation) is not accurate at low SNR and/or when the number of observation samples becomes too small [33]. The large gap between the Cramer-Rao bound and the mean- square estimation error of practical frequency-offset estimators is the result of the estimators sporadically making large errors, referred to as outliers. To analyze this phenomenon, we also discuss the Barankin bound (BB) [27-29] and the Ziv-Zakai bound (ZZB) [20] for frequency-offset estimation. These are more-complicated bounds to compute, but they are considerably tighter than the Cramer-Rao bound at low SNR.

2. Problem Formulation

Throughout the paper, the following signal model is considered:

$$r(n) = a(n) e^{2\pi j(\varphi_0 + \varphi_1 n)} + w(n), \qquad (1)$$

for $n = k_0, \ldots, k_0 + N - 1$, where:

- $a(n)$ is a priori unknown and is referred to as either "multiplicative noise" or "non-constant amplitude."
- φ_0 and φ_1 are the normalized phase shift (at $n = 0$) and the normalized frequency offset of the received signal, respectively. These parameters are also a priori unknown to the receiver and need to be estimated. The absolute value of k_0 determines the difference (in number of symbol intervals) between the start of the received signal and the time instant at which the phase shift, φ_0, is estimated.
- $w(n)$ is circularly symmetric complex-valued additive white Gaussian noise with zero mean, and variance σ_w^2.

By stacking all the available observations into a row vector, we have

$$\mathbf{r} = \mathbf{a}\mathbf{S}\left(\left[\varphi_0, \varphi_1\right]\right) + \mathbf{w}, \qquad (2)$$

where:

- $\mathbf{w}$ is a Gaussian noise vector with zero mean, $\mathbb{E}\left[\mathbf{w}^{\mathrm{T}}\mathbf{w}\right] = \mathbf{0}_N$, and $\mathbb{E}\left[\mathbf{w}^{\mathrm{H}}\mathbf{w}\right] = \sigma_w^2 \mathbf{I}_N$, where $\mathbf{0}_k$ represents a $k \times k$ null matrix, and $\mathbf{I}_k$ represents a $k \times k$ identity matrix. The superscripts $(\cdot)^{\mathrm{T}}$ and $(\cdot)^{\mathrm{H}}$ stand for the transposition and the conjugate-transposition operators, respectively.

- $\mathbf{S}([\varphi_0,\varphi_1])$ is a diagonal matrix with the *n*th diagonal element given by

$$S(n,n;[\varphi_0,\varphi_1]) = e^{2\pi j(\varphi_0+\varphi_1 n)},$$

such that

$$\mathbf{S}([\varphi_0,\varphi_1])\mathbf{S}^{\mathrm{H}}([\varphi_0,\varphi_1]) = \mathbf{I}_N .$$

The signal model of Equations (1) and (2) is encountered in several application fields. A first example is that of digital bandpass communication over an additive-white-Gaussian-noise channel using linear modulation. In that case, $a(n)$ represents the *n*th data symbol passing through the digital bandpass communication channel. The data symbols result from an information bit sequence that is first channel encoded (for better bit-error protection), and then mapped (for higher-bandwidth utilization) to a block of complex numbers belonging to a discrete set Ω, referred to as the symbol constellation. In the digital-communications case, we will also consider that $\sigma_w^2 = N_0/E_s$, with N_0 and E_s assumed to be known. Here, N_0 denotes the noise-power spectral density, and E_s is the symbol energy. The ratio E_s/N_0 is an important measure of the signal quality at the receiver, and is commonly referred to as the *signal-to-noise ratio* (SNR).

As already stated, other application fields where the signal model of Equation (1) can be encountered are that of direction-of-arrival estimation and Doppler radar. In direction-of-arrival estimation, $a(n)$ represents the spatial distribution of the source. In Doppler radar, $a(n)$ represents the Doppler spread of the reference signal. In both cases, it is standard to model the non-constant amplitude as a Gaussian process [1, 3, 34]. Even in digital communications, the process $a(n)$ can sometimes be viewed as a Gaussian process: indeed, in a flat fading channel, $a(n)$ can be the product between a transmitted symbol and a non-constant complex amplitude related to the channel quality. Due to the various scatterers, in a non-line-of-sight (NLOS) channel it is usual to consider that non-constant amplitude as a Gaussian process, and there for to consider its magnitude as a Rayleigh process. It is therefore also referred to as the Rayleigh channel (e.g., [35]).

For the sake of completeness, we note that Equation (1) is only approximate and, in particular, valid only when $|\varphi_1| \ll 1$ [15].

From the observation samples $\{r(n)\}$ in Equation (1), we now want to recover the value of a deterministic parameter vector $\mathbf{u}$ with components u_0, u_1, ... This vector contains (but is not restricted to) the unknown phase shift, φ_0, and the frequency offset, φ_1. A common approach to evaluate the quality of an unbiased estimator for $\mathbf{u}$ consists in comparing its resulting mean-square estimation error with a Cramer-Rao bound, or some other tight, fundamental lower bound on the achievable mean-square estimation error.

3. Deriving the Cramer-Rao Bound

The Cramer-Rao bound results from the inequality $\mathbf{R_u} - \mathbf{J}^{-1} \geq \mathbf{0}$ [32]. Here, $\mathbf{R_u}$ is the error correlation matrix related to the estimation of a deterministic parameter vector $\mathbf{u}$. The notation $\mathbf{A} \geq \mathbf{0}$ indicates that $\mathbf{A}$ is a positive semi-definite matrix, and $\mathbf{J}^{-1}$ denotes the inverse of the Fisher information matrix (FIM), $\mathbf{J}$. The elements of **J** are given by

$$J_{u_k,u_l} = \mathbb{E}\left[\ell_k(\mathbf{u};\mathbf{r})\,\ell_l(\mathbf{u};\mathbf{r})\right], \qquad (3)$$

where J_{u_k,u_l} corresponds to the joint Fisher information for the parameters (u_k,u_l), where $\mathbb{E}[\cdot]$ denotes averaging with respect to $p(\mathbf{r}\,|\,\mathbf{u})$, and where

$$\ell_k(\mathbf{u};\mathbf{r}) = \frac{\partial \ln p(\mathbf{r}|\mathbf{u})}{\partial u_k}$$

is a shorthand notation for the derivative of $\ln p(\mathbf{r}|\mathbf{u})$ with respect to the *k*th parameter, u_k, of $\mathbf{u}$. It easily follows from $\mathbf{R_u} - \mathbf{J}^{-1} \geq \mathbf{0}$ that

$$\mathbb{E}\left[(u_k - \hat{u}_k)^2\right] \geq \mathrm{CRB}(u_k), \qquad (4)$$

where $\mathrm{CRB}(u_k)$ is the *k*th diagonal element of the inverse of the Fisher information matrix, **J**. The right-hand side of the above expression is referred to as the Cramer-Rao bound.

3.1 Non-Constant Complex Amplitude = Digital Data Symbol

In this section we derive the exact Cramer-Rao bound – or, equivalently, the exact Fisher information matrix – for the deterministic parameter vector $\mathbf{u} = [u_0,u_1] = [\varphi_0,\varphi_1]$ from *N* samples of a received linearly modulated digital communication signal in additive white Gaussian noise. We recall that we consider the signal model given by Equations (1) and (2). As is usually done in digital communications, we model the symbol vector, **a,** as a discrete random vector with the following uniform a priori distribution:

$$\Pr\left[\mathbf{a}=\tilde{\mathbf{a}}\right]=\begin{cases}2^{-N_b}, & \tilde{\mathbf{a}}\in\mathcal{S}_0\\ 0, & \tilde{\mathbf{a}}\in\mathcal{S}\setminus\mathcal{S}_0\end{cases}. \qquad (5)$$

Here, $\mathcal{S}$ denotes the set of all possible vectors of N symbols taking values in the symbol constellation set Ω, and $\mathcal{S}_0\subset\mathcal{S}$ denotes the subset of these vectors that result from encoding and mapping an information bit sequence. The distribution of Equation (5) reflects that a one-to-one correspondence exists between the set of all possible sequences of N_b information bits and the data symbol vectors in S_0, while the receiver has no prior knowledge about the transmitted information bit sequence. It is further standard to assume that $\mathbb{E}\left[\mathbf{a}\right]=0$ and that $\mathbb{E}\left[\mathbf{a}^{\mathrm{H}}\mathbf{a}\right]=\mathbf{I}_N$. This assumption holds true for transmissions without channel encoding, and is approximately valid for most practical coded-modulation schemes [36].

A brute-force numerical evaluation of the Fisher information matrix related to the estimation of $\mathbf{u}$ involves replacing the statistical average $\mathbb{E}\left[\cdot\right]$ in Equation (3) by an arithmetical average over a large number of realizations of $\mathbf{r}$ that are computer-generated according to the conditional distribution $p\left(\mathbf{r}\mid\mathbf{u}\right)$. The numerical evaluation of the Fisher information matrix further requires the computation of the derivatives $\ell_k\left(\mathbf{u};\mathbf{r}\right)$, $k=0,1$, which correspond to the realizations of $\mathbf{r}$ given $\mathbf{u}$. These derivatives can be put into the following form [37]:

$$\ell_k\left(\mathbf{u};\mathbf{r}\right)=\sum_{\tilde{\mathbf{a}}}\frac{\partial\ln p\left(\mathbf{r}\middle|\mathbf{a}=\tilde{\mathbf{a}},\mathbf{u}\right)}{\partial u_k}\Pr\left[\mathbf{a}=\tilde{\mathbf{a}}\middle|\mathbf{r},\mathbf{u}\right]. \qquad (6)$$

As $\mathbf{p}\left(\mathbf{r}\middle|\mathbf{a},\mathbf{u}\right)$ is Gaussian, the logarithm $\ln p\left(\mathbf{r}\middle|\mathbf{a},\mathbf{u}\right)$ is readily available in closed form:

$$\ln p\left(\mathbf{r}\middle|\mathbf{a},\mathbf{u}\right)\propto-\frac{E_s}{N_0}\left|\mathbf{r}-\mathbf{a}\mathbf{S}\left(\mathbf{u}\right)\right|^2.$$

Differentiating with respect to u_k yields

$$\frac{\partial\ln p\left(\mathbf{r}\middle|\mathbf{a},\mathbf{u}\right)}{\partial u_k}=-\frac{2E_s}{N_0}\Re\left\{\left[\mathbf{r}-\mathbf{a}\mathbf{S}\left(\mathbf{u}\right)\right]^{\mathrm{H}}\left[\mathbf{a}\frac{\partial\mathbf{S}\left(\mathbf{u}\right)}{\partial u_k}\right]\right\}$$

The joint symbol a posteriori probabilities (APP) $\Pr\left[\mathbf{a}\middle|\mathbf{r},\mathbf{u}\right]$ in Equation (6) can be computed from $p\left(\mathbf{r}\middle|\mathbf{a},\mathbf{u}\right)$ and $\Pr\left[\mathbf{a}\right]$ according to

$$\Pr\left[\mathbf{a}\middle|\mathbf{r},\mathbf{u}\right]=\frac{\Pr\left[\mathbf{a}\right]p\left(\mathbf{r}\middle|\mathbf{a},\mathbf{u}\right)}{\sum_{\tilde{\mathbf{a}}}p\left(\mathbf{r}\middle|\mathbf{a}=\tilde{\mathbf{a}},\mathbf{u}\right)\Pr\left[\mathbf{a}=\tilde{\mathbf{a}}\right]}. \qquad (7)$$

Although this procedure yields the exact derivatives $\ell_k\left(\mathbf{u};\mathbf{r}\right)$, $k=0,1$, the summations in Equations (6) and (7) give rise to a computational complexity that is exponential in the burst size, N.

It was shown in [16] and [37] that the computational complexity associated with the evaluation of the Cramer-Rao bound can be drastically reduced by taking into account the specific (linearly modulated) structure of the useful signal in Equation (1).

Because $\mathbf{S}\left(\mathbf{u}\right)\mathbf{S}^{\mathrm{H}}\left(\mathbf{u}\right)=\mathbf{I}_N$ does not depend on $\mathbf{u}$, we obtain

$$\frac{\partial\ln p\left(\mathbf{r}\middle|\mathbf{a},\mathbf{u}\right)}{\partial u_k}=\frac{2E_s}{N_0}\Re\left\{\mathbf{r}\left[\frac{\partial\mathbf{S}\left(\mathbf{u}\right)}{\partial u_k}\right]^{\mathrm{H}}\mathbf{a}^{\mathrm{H}}\right\}.$$

Substituting the above expression into Equation (6) then yields

$$\ell_k\left(\mathbf{u};\mathbf{r}\right)=\frac{2E_s}{N_0}\Re\left\{\mathbf{r}\left(\frac{\partial\mathbf{S}\left(\mathbf{u}\right)}{\partial u_k}\right)^{\mathrm{H}}\boldsymbol{\mu}^{\mathrm{H}}\left(\mathbf{r},\mathbf{u}\right)\right\}, \qquad (8)$$

where $\boldsymbol{\mu}\left(\mathbf{r},\mathbf{u}\right)$ is a shorthand notation for the a posteriori average of the symbol vector $\mathbf{a}$ with N components:

$$\mu\left(n;\mathbf{r},\mathbf{u}\right)=\mathbb{E}_{\mathbf{a}}\left[a\left(n\right)\middle|\mathbf{r},\mathbf{u}\right] \qquad (9)$$

$$=\sum_{\omega\in\Omega}\omega\Pr\left[a\left(n\right)=\omega\middle|\mathbf{r},\mathbf{u}\right], \qquad (10)$$

where Ω denotes the set of constellation points, and the averaging $\mathbb{E}_{\mathbf{a}}\left[\cdot\middle|\mathbf{r},\mathbf{u}\right]$ is with respect to $\Pr\left[\mathbf{a}\middle|\mathbf{r},\mathbf{u}\right]$. We emphasize that no approximation is involved in obtaining Equation (10): the right-hand side simply expresses the a posteriori average of the nth data symbol $a\left(n\right)$ in terms of the *marginal* a posteriori probabilities of $a\left(n\right)$, rather than the *joint* a posteriori probabilities of all components of $\mathbf{a}$.

Computing the marginal a posteriori probabilities from the joint a posteriori probabilities still requires a complexity that increases exponentially with N. However, in most practical scenarios, the required marginal symbol a posteriori probabilities can be directly computed in an efficient way by applying the sum-product algorithm to a factor graph (FG) representing a suitable factorization of the joint-symbol a posteriori probabilities [38]. The application of the sum-product algorithm on a graph that corresponds to a tree (i.e., a cycle-free factor graph) is straightforward, and yields the exact marginals. When the graph contains cycles, the sum-product algorithm becomes an iterative procedure that yields only an approximation of the marginals after convergence. However, when the cycles in the graph are large, the resulting marginals turn out to be quite accurate. When using this factor-graph-based approximation technique to compute the required marginal symbol a posteriori probabilities, computing the derivatives $\ell_k(\mathbf{u};\mathbf{r})$, $k=0,1$ according to Equation (8) for a given realization of **r** given **u** yields a complexity that is linear (and not exponential) in the number of data symbols *N*. The above expression and evaluation procedure is the main result for the Cramer-Rao bound derivations. It allows a fast evaluation of the Cramer-Rao bound, and holds for any channel code and any symbol constellation.

For specific hypotheses about the channel code and the symbol constellation, the complexity associated with evaluating the Fisher information matrix, or, equivalently, the Cramer-Rao bounds, can be further reduced (see, e.g., [12-13, 15, 19]). We mention a result from [15] for arbitrarily mapped uncoded linear modulation. In that case, all transmitted data symbols are statistically independent and equi-probable, such that the a priori distribution of **a**, Equation (5), reduces to

$$\Pr\left[\mathbf{a}=\tilde{\mathbf{a}}\right]=2^{-N_b},\ \forall\tilde{\mathbf{a}}\in\mathcal{S}. \tag{11}$$

Taking into account Equation (11), it is easily verified from Equations (9), (10), and (7) that the components of the a posteriori average of the data-symbol vector **a** reduce to

$$\mu\left[n;r(n),\mathbf{u}\right]=\frac{\sum_{\omega\in\Omega}\omega e^{\left\{\frac{E_s}{N_0}\left[2\Re\left\{r(n)e^{-2\pi j(\varphi_0+\varphi_1 n)}\omega^*\right\}-|\omega|^2\right]\right\}}}{\sum_{\omega\in\Omega}e^{\left\{\frac{E_s}{N_0}\left[2\Re\left\{r(n)e^{-2\pi j(\varphi_0+\varphi_1 n)}\omega^*\right\}-|\omega|^2\right]\right\}}}, \tag{12}$$

which only depend on **r** through $r(n)$. Taking this into account, it was shown in [15] that with *N* odd-valued and $k_0=-\frac{1}{2}(N-1)$ (i.e., φ_0 is the phase shift at the burst center), the Cramer-Rao bounds can be written in the following form:

$$\left[\mathrm{CRB}(\varphi_0)\right]^{-1}=8\pi^2\frac{E_s}{N_0}NR_\Omega\left(\frac{E_s}{N_0}\right), \tag{13}$$

$$\left[\mathrm{CRB}(\varphi_1)\right]^{-1}=\left[\mathrm{CRB}(\varphi_0)\right]^{-1}\frac{\left(N^2-1\right)}{12}, \tag{14}$$

where

$$R_\Omega\left(\frac{E_s}{N_0}\right)=\frac{2E_s}{N_0}\mathbb{E}\left[\Im\left\{\mu^*\left[n;r(n),\mathbf{u}\right]r(n)S^*(n,n;\mathbf{u})\right\}^2\right] \tag{15}$$

It was further shown in [15] that $J_{\varphi_0,\varphi_1}=J_{\varphi_1,\varphi_0}=0$. This means that the estimation of the phase shift, φ_0, at the center of the observation interval is independent of the frequency (φ_1) estimation problem. We observe that $\mathrm{CRB}(\varphi_0)$ is inversely proportional to the number of available signal samples, *N*, whereas $\mathrm{CRB}(\varphi_0)$ is inversely proportional to $N\left(N^2-1\right)\approx N^3$, where the approximation holds for large *N*. We further observe that $\mathrm{CRB}(\varphi_0)$ and $\mathrm{CRB}(\varphi_1)$ are proportional to the same factor $R_\Omega\left(\frac{E_s}{N_0}\right)$ that depends on the symbol constellation and on the SNR, but not on the number of available signal samples. The numerical evaluation of the Cramer-Rao bounds from Equations (13) and (14) involves replacing the statistical average $\mathbb{E}[\cdot]$ in Equation (15) by an arithmetical average over a large number of realizations of $r(n)S^*(n,n;\mathbf{u})$. This procedure is significantly less complex than the evaluation of the Fisher information matrix entries according to Equations (3) and (8)-(10) using the factor-graph-approach, because the a posteriori symbol average, Equation (12), is available in closed form, and the average that needs to be computed in Equation (15) is with respect to a complex-valued *scalar*, rather than a complex-valued vector of size *N*.

In spite of all the efforts made in the literature with respect to computing the Fisher information matrix for linear modulation, an explicit analytical closed-form expression for the Cramer-Rao bounds still does not exist. The main contributions of the research conducted in [12-13] and [15-19] was in the derivation of new procedures that allow a more-efficient (hence, faster) numerical evaluation of the Cramer-Rao bounds. Unfortunately, the expressions that lead to (and/or come as by-products of) these evaluation procedures usually don't bring much insight into the behavior of the Fisher information matrix (e.g., as a function of the parameters that describe the coded modulation scheme).

We will see in the simulation section of the paper that

- For a given symbol constellation set, Cramer-Rao bounds for coded and uncoded transmission are equal at sufficiently high SNR. However, at lower SNR the Cramer-Rao bound for coded transmission is significantly lower than the Cramer-Rao bound for uncoded transmission.

- For a given channel code, the Cramer-Rao bounds increase when the minimum Euclidean distance between the constellation points decreases.

To avoid the computational complexity associated with the evaluation of the true Cramer-Rao bounds, asymptotic Cramer-Rao bounds (ACRBs) were considered in [14] and [11] for the case of uncoded linear modulation. This resulted in closed-form analytical expressions for the Cramer-Rao bound that only hold for sufficiently low or high SNR. The high-SNR asymptotic Cramer-Rao bounds are shown to coincide with the modified Cramer-Rao bound (MCRB). This is another lower bound on the mean square estimation error of any unbiased estimator, which is simpler to evaluate but looser than the exact (true) Cramer-Rao bound. We will come back to the modified Cramer-Rao bound later in this paper. For the low-SNR asymptotic Cramer-Rao bounds, the following expressions were presented in [14], assuming N odd valued and $k_0 = -\frac{1}{2}(N-1)$

$$\left[\mathrm{ACRB}(\varphi_0)\right]^{-1}_{|\mathrm{lowSNR}} = 8\pi^2 \left(\frac{E_s}{N_0}\right)^L N \frac{L^2 |f_L|^2}{L!},$$

$$\left[\mathrm{ACRB}(\varphi_1)\right]^{-1}_{|\mathrm{lowSNR}} = \left[\mathrm{ACRB}(\varphi_0)\right]^{-1}_{|\mathrm{lowSNR}} \frac{\left(N^2-1\right)}{12}, \tag{16}$$

where L is related to the symmetry angle $\frac{2\pi}{L}$ of the constellation, and $f_L = \mathbb{E}\left\{\left[a(k)\right]^L\right\}$. We observe that at sufficiently low SNR, the Cramer-Rao bounds are determined by the symmetry angle of the constellation, and evolve in inverse proportion to the Lth power of the SNR.

3.2 Non-Constant Complex Amplitude = A Gaussian Process

We recall that we consider the signal model given by Equation (1). In the rest of this section, we just assume that $k_0 = 0$. However, in this section we add extra assumptions on the non-constant amplitude $a(n)$. As usually done in Doppler radar, direction-of-arrival estimation, or digital communication over a Rayleigh flat-fading channel, the non-constant amplitude, $a(n)$, is assumed to be a zero-mean Gaussian stationary process with correlation $c_a(\tau) = \mathbb{E}\left[a(n+\tau)a^*(n)\right]$ and pseudo-correlation $p_a(\tau) = \mathbb{E}\left[a(n+\tau)a(n)\right]$. The spectrum and pseudo-spectrum are respectively denoted as follows:

$$C_a\left(e^{2i\pi f}\right) = \sum_{\tau\in\mathbb{Z}} c_a(\tau) e^{-2i\pi f\tau}$$

and

$$P_a\left(e^{2i\pi f}\right) = \sum_{\tau\in\mathbb{Z}} p_a(\tau) e^{-2i\pi f\tau}.$$

By construction, one can remark that $P_a\left(e^{2i\pi f}\right) = P_a\left(e^{-2i\pi f}\right)$ Moreover, the entire statistics $\left\{c_a(\tau), p_a(\tau)\right\}_{\tau\in\mathbb{Z}}$ of $a(n)$ only depend on a finite number, K, of real-valued unknown parameters, denoted by $\{\alpha_k\}_{k=1,\ldots,K}$. The non-constant amplitude process, $a(n)$, can be real-valued or complex-valued. In the case of a complex-valued process, $a(n)$ can further be *circular* (which means the process distribution is insensitive to any rotation, and thus means that $\mathbb{E}\left[a(n)a(n+\tau)\right] = 0$ for all τ) or *noncircular* (there exists at least one τ_0 such that $\mathbb{E}\left[a(n)a(n+\tau_0)\right] \neq 0$). One can notice that a real-valued process is by definition noncircular. Based on the Cramer-Rao bound, we will see hereafter that the estimation quality can be split into two classes in regard to the circularity/non-circularity property of the process. In contrast, the estimation performance is independent of the nature of the process' values (real or complex). For further information on the non-circularity property, the reader may refer to [15, 39].

Below, we will first derive the Fisher information matrix when the number of available samples, N, is finite (i.e., the non-asymptotic case). Since once again the expression obtained for the Fisher information matrix does not provide additional insights, it is of great interest to further simplify the Fisher information matrix expression by also considering the case for N going to infinity (i.e., the asymptotic case). The resulting Cramer-Rao bounds are referred to as Gaussian Cramer-Rao bounds (GCRB).

3.2.1 Non-Asymptotic Case

We next derive the exact Gaussian Cramer-Rao bound, or, equivalently, the exact Gaussian Fisher information matrix, $\mathbf{J}$, for the deterministic parameter vector $\mathbf{u} = \left[\varphi_0, \varphi_1, \sigma_w^2, \alpha_1, \ldots, \alpha_K\right]$, when N samples of $r(n)$ are available. In order to use well-known results on the Fisher information matrix [40], we work with real-valued processes. We consider $\breve{\mathbf{r}} = \left\{\Re[\mathbf{r}], \Im[\mathbf{r}]\right\}$, which is a multivariate Gaussian variable with zero mean and covariance matrix $\breve{\mathbf{C}}_\mathbf{r}$.

The Fisher information matrix for a multivariate Gaussian observation vector $\breve{\mathbf{r}}$ has a special form. As one can check that $\breve{\mathbf{C}}_\mathbf{r}$ is symmetric, formula (5.2.1) in [40] holds, and this leads to

$$J_{u_k,u_l} = \frac{1}{2}\mathrm{Tr}\left(\frac{\partial \breve{\mathbf{C}}_\mathbf{r}}{\partial u_k} \breve{\mathbf{C}}_\mathbf{r}^{-1} \frac{\partial \breve{\mathbf{C}}_\mathbf{r}}{\partial u_l} \breve{\mathbf{C}}_\mathbf{r}^{-1}\right),$$

where $\mathrm{Tr}(\cdot)$ is the trace operator.

After straightforward algebraic manipulations, we can show that

$$J_{u_k,u_l} = \frac{1}{2}\mathrm{Tr}\left(\frac{\partial \tilde{\mathbf{C}}_\mathbf{r}}{\partial u_k} \tilde{\mathbf{C}}_\mathbf{r}^{-1} \frac{\partial \tilde{\mathbf{C}}_\mathbf{r}}{\partial u_l} \tilde{\mathbf{C}}_\mathbf{r}^{-1}\right),$$

where $\tilde{\mathbf{C}}_\mathbf{r}$ is the covariance matrix of the random vector $\tilde{\mathbf{r}} = \left[\mathbf{r}, \mathbf{r}^*\right]$, and takes the following form:

$$\tilde{\mathbf{C}}_\mathbf{r} = \begin{bmatrix} \mathbf{C}_\mathbf{r} & \mathbf{P}_\mathbf{r} \\ \mathbf{P}_\mathbf{r}^* & \mathbf{C}_\mathbf{r}^* \end{bmatrix}$$

with $\mathbf{C}_\mathbf{r} = \mathbf{E}\left[\mathbf{r}^\mathrm{H}\mathbf{r}\right]$ and $\mathbf{P}_\mathbf{r} = \mathbf{E}\left[\mathbf{r}^\mathrm{T}\mathbf{r}\right]$.

One can remark that $\tilde{\mathbf{C}}_\mathbf{r} = \tilde{\mathbf{S}}\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)\tilde{\mathbf{S}}^\mathrm{H}$, where $\tilde{\mathbf{S}} = \left\{\mathbf{S}\left[(\varphi_0, \varphi_1)\right], \mathbf{0}_N; \mathbf{0}_N, \mathbf{S}^*\left[(\varphi_0, \varphi_1)\right]\right\}$, and where $\tilde{\mathbf{C}}_\mathbf{a}$ is defined in a similar way as $\tilde{\mathbf{C}}_\mathbf{r}$. As $\tilde{\mathbf{C}}_\mathbf{a}$ does not depend on the phase parameters, we obtain the following expressions for the Fisher information matrix:

$$J_{\alpha_k,\alpha_l} = \frac{1}{2}\mathrm{Tr}\left[\frac{\partial \tilde{\mathbf{C}}_\mathbf{a}}{\partial \alpha_k}\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)^{-1}\frac{\partial \tilde{\mathbf{C}}_\mathbf{a}}{\partial \alpha_l}\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)^{-1}\right],$$

$$J_{\sigma_w^2,\sigma_w^2} = \frac{1}{2}\mathrm{Tr}\left[\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)^{-2}\right],$$

$$J_{\alpha_k,\sigma_w^2} = \frac{1}{2}\mathrm{Tr}\left[\frac{\partial \tilde{\mathbf{C}}_\mathbf{a}}{\partial a_k}\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)^{-2}\right],$$

$$J_{\varphi_k,\varphi_l} = 2\pi^2 \mathrm{Tr}\left[\mathbf{D}_k\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)\mathbf{D}_l\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)^{-1}\right.$$

$$+\mathbf{D}_l\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)\mathbf{D}_k\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)^{-1}$$

$$\left. -2\mathbf{D}_k\mathbf{D}_l\right]$$

$$J_{\alpha_k,\varphi_k} = i\pi \mathrm{Tr}\left\{\frac{\partial \tilde{\mathbf{R}}_\mathbf{a}}{\partial \alpha_k}\left[\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)^{-1}\mathbf{D}_k\right.\right.$$

$$\left.\left. -\mathbf{D}_k\left(\tilde{\mathbf{C}}_\mathbf{a} + \sigma_w^2 \mathbf{I}_{2N}\right)^{-1}\right]\right\}$$

$$J_{\sigma_w^2,\varphi_k} = 0,$$

where $\mathbf{D}_k = \left[\mathbf{d}_k, \mathbf{0}_N; \mathbf{0}_N, -\mathbf{d}_k\right]$ for $k = 0,1$ with $\mathbf{d}_0 = \mathbf{I}_N$ and $\mathbf{d}_1 = \mathrm{diag}\left[\left(0, \cdots, N-1\right)\right]$. The above expressions were given in [31], and partially in [3] when $a(n)$ was circular and complex-valued. When $a(n)$ is circular and complex-valued, the term $J_{\varphi_0,\varphi_0} = 0$, which means that the constant phase is not identifiable when the pseudo-correlation is zero. Consequently, only a non-null pseudo-correlation enables us to estimate the constant phase. Apart from this comment about the constant phase, it is difficult to provide more insights with these expressions, and to distinguish the difference between the circular case and the noncircular case. We therefore now move on to the asymptotic case, i.e., for N sufficiently large.

3.2.2 Asymptotic Case

When N becomes large, we have to separately treat the circular case and the noncircular case. Let us begin with the circular case.

3.2.2.1 Circular Case

When the signal, $r(n)$, is circular, one can remark that $r(n)$ is stationary, due to our signal model given in Equation (1). This enables us to simplify the asymptotic expressions for the Fisher information matrix by applying Whittle's formula [41].

In [3], the asymptotic expressions for the Cramer-Rao bound are given for $C_a(\tau)$ real valued and positive. The latter assumption has been justified by many other authors [1, 23-25]. For instance, if $a(n)$ is associated with the Doppler-spread phenomenon, $C_a(\tau)$ often follows the Jakes model [42, 43], and thus $C_a(\tau) = \sigma_a^2 J_0(\Delta\tau)$, where σ_a^2 is the variance of $a(n)$, Δ is the Doppler spread, and $J_0(\cdot)$ is the Bessel function of first kind. In such a case, one can prove that the estimates of $[\varphi_0, \varphi_1]$ are decoupled from the other parameters, $\left[\sigma_w^2, \alpha_1, \cdots, \alpha_K\right]$. As remarked in the previous subsection, the phase, φ_0, cannot be estimated in the circular case. As a consequence, we can focus only on J_{φ_1,φ_1}. After tedious algebraic manipulations, one can find that

$$\lim_{N\to\infty}\frac{1}{N}J_{\varphi_1,\varphi_1} = \delta$$

with

$$\delta = \int_0^1\left[\frac{C_a'\left(e^{2i\pi f}\right)}{C_a\left(e^{2i\pi f}\right) + \sigma_w^2}\right]^2 df$$

and where $C_a'\left(e^{2i\pi f}\right)$ is the derivative function of $C_a\left(e^{2i\pi f}\right)$ with respect to f. As the Cramer-Rao bound is the inverse of the Fisher information matrix, we have

$$\mathrm{GCRB}(\varphi_1) \approx \frac{1}{\delta N} \quad \text{(circular case)}.$$

We remark that the frequency can be estimated as soon as $\delta \neq 0$, i.e., as soon as the process $a(n)$ does not have a flat spectrum. We thus need to have a colored Gaussian non-constant amplitude process – and not a white Gaussian non-constant amplitude process – to be able to estimate the frequency, if the process is circular. Moreover, the Cramer-Rao bound is proportional to $1/N$, and so the minimum-achievable mean square estimation error decreases quite slowly with respect to the number of available samples.

3.2.2.2 Non-Circular Case

Unlike [3], here we cannot apply Whittle's formula [41], because $r(n)$ is not stationary with respect to its pseudo-correlation. Below, the results introduced are in fact obtained via theorems dealing with the inversion of (large) Toeplitz matrices [44, 45].

After simple but tedious calculations, the Fisher information matrix was found in [31] to be

$$\lim_{N\to\infty} \frac{1}{N} J_{\alpha_k,\alpha_l} = \frac{1}{2}\theta_{k,l},$$

$$\lim_{N\to\infty} \frac{1}{N} J_{\sigma_w^2,\sigma_w^2} = \frac{1}{2}\gamma,$$

$$\lim_{N\to\infty} \frac{1}{N} J_{\alpha_k,\sigma_w^2} = \frac{1}{2}\beta_k,$$

$$\lim_{N\to\infty} \frac{1}{N} J_{\varphi_0,\varphi_0} = 16\pi^2\xi,$$

$$\lim_{N\to\infty} \frac{1}{N^3} J_{\varphi_1,\varphi_1} = \frac{16\pi^2}{3}\xi,$$

$$\lim_{N\to\infty} \frac{1}{N^2} J_{\varphi_0,\varphi_1} = 8\pi^2\xi,$$

$$\lim_{N\to\infty} \frac{1}{N} J_{\alpha_k,\varphi_0} = 4\pi\mu_k,$$

$$\lim_{N\to\infty} \frac{1}{N^2} J_{\alpha_k,\varphi_1} = 2\pi\mu_k,$$

where

$$\theta_{k,l} = \int_0^1 \frac{1}{X\left(e^{2i\pi f}\right)^2} \frac{\partial X\left(e^{2i\pi f}\right)}{\partial \alpha_k} \frac{\partial X\left(e^{2i\pi f}\right)}{\partial \alpha_l} df$$

$$+\int_0^1 \frac{1}{X\left(e^{2i\pi f}\right)} \left[Q_{k,l}^{(P_a)}\left(e^{2i\pi f}\right) - Q_{k,l}^{(C_a+\sigma_w^2)}\left(e^{2i\pi f}\right)\right] df$$

$$\gamma = \int_0^1 \frac{1}{X\left(e^{2i\pi f}\right)^2} \left\{ \left[C_a\left(e^{2i\pi f}\right) + \sigma_w^2 \right]^2 \right.$$

$$\left. + \left[\underline{C_a}\left(e^{2i\pi f}\right) + \sigma_w^2 \right]^2 + 2\underline{P_a}\left(e^{2i\pi f}\right) P_a\left(e^{2i\pi f}\right) \right\} df$$

$$\beta_k = \int_0^1 \frac{1}{X\left(e^{2i\pi f}\right)} \frac{\partial X\left(e^{2i\pi f}\right)}{\partial \alpha_k} df$$

$$\mu_k = \Im m\left[\int_0^1 \frac{\underline{P_a}\left(e^{2i\pi f}\right)}{X\left(e^{2i\pi f}\right)} \frac{\partial P_a\left(e^{2i\pi f}\right)}{\partial \alpha_k} df \right]$$

$$\xi = \int_0^1 \frac{P_a\left(e^{2i\pi f}\right) \underline{P_a}\left(e^{2i\pi f}\right)}{X\left(e^{2i\pi f}\right)} df$$

with

$$\underline{\nu}\left(e^{2i\pi f}\right) = \overline{\nu\left(e^{-2i\pi f}\right)},$$

$$Q_{k,l}^{(\nu)}\left(e^{2i\pi f}\right)$$

$$= \frac{\partial \nu\left(e^{2i\pi f}\right)}{\partial \alpha_k} \frac{\partial \underline{\nu}\left(e^{2i\pi f}\right)}{\partial \alpha_l} + \frac{\partial \underline{\nu}\left(e^{2i\pi f}\right)}{\partial \alpha_k} \frac{\partial \nu\left(e^{2i\pi f}\right)}{\partial \alpha_l}$$

$$X\left(e^{2i\pi f}\right) = \left[C_a\left(e^{2i\pi f}\right) + \sigma_w^2\right]\left[\underline{C_a}\left(e^{2i\pi f}\right) + \sigma_w^2\right]$$

$$- P_a\left(e^{2i\pi f}\right)\underline{P_a}\left(e^{2i\pi f}\right)$$

We next study different scenarios. First, we consider the case where the receiver knows $\left[\sigma_w^2, \alpha_1, \cdots, \alpha_K\right]$, i.e., the statistics of multiplicative and additive noise. In this case, the Gaussian Cramer-Rao bounds result from the inverse of the 2×2 Fisher information matrix

$$\mathbf{J}_{[\varphi_0,\varphi_1]} = \begin{bmatrix} J_{\varphi_0,\varphi_0} & J_{\varphi_0,\varphi_1} \\ J_{\varphi_0,\varphi_1} & J_{\varphi_1,\varphi_1} \end{bmatrix}.$$

This yields

$$\mathrm{GCRB}\left(\varphi_0\right)_{|\text{noise statistics known}} = \frac{1}{4\pi^2 \xi N}$$

and

$$\mathrm{GCRB}\left(\varphi_1\right)_{|\text{noise statistics known}} = \frac{3}{4\pi^2 \xi N^3}.$$

Second, in the case where $\left[\sigma_w^2, \alpha_1, \cdots, \alpha_K\right]$ are unknown at the receiver, we obtain (see [31])

$$\mathrm{GCRB}\left(\varphi_0\right)_{|\text{noise statistics unknown}}$$

$$= \mathrm{GCRB}\left(\varphi_0\right)_{|\text{noise statistics known}} + \frac{m}{16\pi^2 \xi^2 N}.$$

Here, m is a bounded scalar taking the following form:

$$m = \boldsymbol{\mu}^{\mathrm{T}}\left[\boldsymbol{\theta}/2 - \boldsymbol{\mu}\boldsymbol{\mu}^{\mathrm{T}}/\xi - \boldsymbol{\beta}\boldsymbol{\beta}^{\mathrm{T}}/(2\gamma)\right]^{-1}\boldsymbol{\mu}$$

where

$$\boldsymbol{\theta} = \left(\theta_{k,l}\right)_{1 \le k, l \le K},$$

$$\boldsymbol{\beta} = \left(\beta_k\right)_{1 \le k \le K},$$

$$\boldsymbol{\mu} = \left(\mu_k\right)_{1 \le k \le K}.$$

Using the previous expressions for the asymptotic Cramer-Rao bound, we make the following comments:

- The convergence rates for the phase and frequency estimations are $1/N$ and $1/N^3$, respectively, regardless of the color of the multiplicative noise. Recall that for circular complex-valued processes, the phase is not identifiable, and the frequency is identifiable only if the multiplicative noise is colored, with a convergence rate of $1/N$. Notice that a real-valued process can be viewed as a specific case of a noncircular complex-valued process where the imaginary part is zero. Consequently, in terms of performance, the main cutoff is not complex/real, but circular/noncircular.

- We recall that the Cramer-Rao bound associated with the "pure" frequency-estimation issue (i.e., only disturbed by a constant amplitude) is proportional to $1/N^3$ [46]. Consequently, thanks to the non-circularity property of the non-constant amplitude, the non-constant amplitude does not lead to a significant loss in performance.

- Surprisingly, the same frequency-estimation performance is obtained whether the statistics of $a(n)$ are known or not.

- The frequency-estimation performance depends only on ξ, which refers to an information rate provided by the non-circularity. Indeed, the performance improves when ξ increases.

- In the noiseless case, we observe a floor effect (i.e., $\mathrm{CRB}\left(\varphi_1\right) \neq 0$ when $\sigma_w^2 = 0$). This effect vanishes if $C_a\left(e^{2i\pi f}\right)C_a\left(e^{-2i\pi f}\right) = P_a\left(e^{2i\pi f}\right)P_a\left(e^{-2i\pi f}\right)$. This condition is fulfilled, for example, when the multiplicative noise is real-valued.

As a conclusion, we remind the reader that the Gaussian Cramer-Rao bound is of interest in many applications: Doppler radar, direction-of-arrival estimation, and digital communication over Rayleigh flat-fading channels. If the non-constant amplitude is non-Gaussian, the Gaussian Cramer-Rao bound is not a lower bound for the estimation problem anymore. Nevertheless, it is still of interest, since the Gaussian Cramer-Rao bound is a lower bound for any second-order-based estimator (well-adapted to, e.g., digital BPSK modulation) [47]. Consequently, the

Gaussian Cramer-Rao indicates what is the best expected performance if we carry out an estimator based only on mean and correlation.

Actually, in the non-Gaussian case, the non-circularity may also play a significant role. For example, if $a(n)$ is assumed to belong to a QAM modulation, $a(n)$ is circular at second order (i.e., $\mathbb{E}\left[a(n)a(n)\right]=0$, but is non-circular at fourth order ($\mathbb{E}\left[a(n)a(n)a(n)a(n)\right]\neq 0$). Thanks to this fourth-order non-circularity, we are able to build an estimator for which the mean-square estimation error is proportional to $1/N^3$, and not $1/N$ [47, 48]. This is not in contradiction with the previous results, since a QAM modulation is not Gaussian, and so the high-order statistics of $a(n)$ strongly help to improve the estimation performance.

4. Deriving the Modified Cramer-Rao Bound

To overcome the complexity concern of deriving the true Cramer-Rao bound[1] in the non-Gaussian case, it is possible to define other lower-Cramer-Rao-bound-like bounds that are easier to compute but less tight than the true bound. The most well-spread is the so-called *modified* Cramer-Rao bound (MCRB) [9, 10]. Once again, we will restrict our analysis to the estimation of $\mathbf{u}=\left[\varphi_0,\varphi_1\right]$. The elements of the *modified* Fisher information matrix (MFIM) are then defined as follows:

$$\mathcal{J}_{\varphi_k,\varphi_l}=\mathbb{E}_{\mathbf{a}}\left[\frac{\partial \ln p\left(\mathbf{r}\,|\,\mathbf{u},\mathbf{a}\right)}{\partial\varphi_k}\frac{\partial \ln p\left(\mathbf{r}\,|\,\mathbf{u},\mathbf{a}\right)}{\partial\varphi_l}\right].$$

After standard algebraic manipulations, we obtain

$$\mathcal{J}_{\varphi_k,\varphi_l}=\frac{8\pi^2}{\sigma_w^2}\mathbb{E}_{\mathbf{a}}\left[\mathbf{a}\mathbf{d}_k\mathbf{d}_l\mathbf{a}^{\mathrm{H}}\right].$$

For large *N*, the resulting modified Cramer-Rao bounds were given in [9, 10]. We have

$$\mathrm{MCRB}\left(\varphi_0\right)\approx\frac{\left[\left(k_0+N-1\right)^3-\left(k_0\right)^3\right]\sigma_w^2}{2\pi^2c_a\left(0\right)N^4}$$

and

$$\mathrm{MCRB}\left(\varphi_1\right)\approx\frac{3\sigma_w^2}{2\pi^2c_a\left(0\right)N^3}.$$

We note that in the case of linear modulation it was assumed that $c_a\left(0\right)=1$, such that the above modified Cramer-Rao bounds, for *N* large and odd valued and $k_0=\left(N-1\right)/2$, reduce to the Cramer-Rao bounds from Equations (13) and (14) upon the factor $R_\Omega\left(\sigma_w^{-2}\right)$ from Equation (15). We have the following comments:

- The derivation of the modified Cramer-Rao bounds is very easy and enables us to obtain simple closed-form expressions.

- These expressions seem to be "too" simple and do not provide a lot of information, since the Cramer-Rao bound does not depend on the nature of $a(n)$ (circular/noncircular in the Gaussian case, channel code and symbol constellation in the non-Gaussian case), while we have seen before that this is crucial information (see the previous discussion in the Gaussian case and the asymptotic Cramer-Rao-bound expressions in non-Gaussian case).

- Nevertheless, the modified Cramer-Rao bound can sometimes be of great interest. Indeed, if $a(n)$ belongs to a finite set of symbol constellation points Ω, the true Cramer-Rao bound (for which no explicit expressions are available) is well approximated by the modified Cramer-Rao bound at high SNR [11].
- An unexpected consequence of the previous remark is the following. Let us consider the modified Cramer-Rao bound for estimating the frequency offset in the case of digital communication using a BPSK symbol constellation set, i.e., $a(n)$ takes values in the set $\{-1,1\}$ which implies that $L=2$ and $f_L=c_a\left(0\right)=1$. We have

$$\mathrm{MCRB}\left(\varphi_1\right)_{\mathrm{BPSK}}\approx\frac{3\sigma_w^2}{2\pi^2N^3}.$$

Due to the previous item, there is equivalence between MCRB and ACRB at high SNR for BPSK. We therefore know that

$$\mathrm{ACRB}\left(\varphi_1\right)_{|\text{high SNR},\mathrm{BPSK}}=\frac{3\sigma_w^2}{2\pi^2N^3}$$

and, thanks to Equation (16), we get

$$\mathrm{ACRB}\left(\varphi_1\right)_{|\text{low SNR},\mathrm{BPSK}}=\frac{3\sigma_w^4}{4\pi^2N^3}.$$

Obviously, at low SNR, the true Cramer-Rao bound for BPSK starts to seriously deviate from the modified Cramer-Rao bound.

If we inspect the Gaussian Cramer-Rao bound for uncorrelated and real-valued Gaussian $a(n)$, we obtain

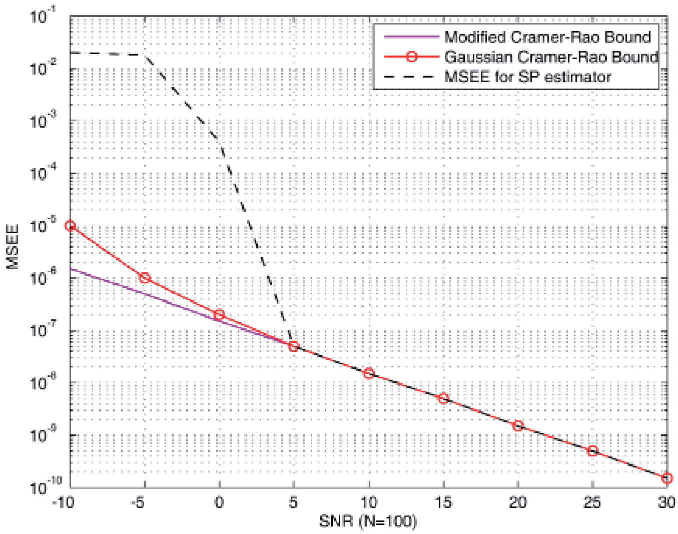

Figure 1. The modified Cramer-Rao bound (MCRB), Gaussian Cramer-Rao bound (GCRB), and mean-square estimation error (MSEE) for the square-power estimator as a function of the SNR.

$$\mathrm{GCRB}(\varphi_1) = \frac{3\left[2\sigma_w^2 + \sigma_w^4\right]}{4\pi^2 N^3}.$$

Surprisingly, the Gaussian Cramer-Rao bound predicts the performance of a BPSK based non-constant amplitude well for both low *and* high SNR, whereas a BPSK constellation is not Gaussian at all! Consequently, the Gaussian Cramer-Rao bound is a powerful tool for analyzing the frequency mean-square estimation error in the BPSK context, whereas the modified Cramer-Rao bound is not (except at high SNR).

5. Deriving the Barankin Bound

Let us reconsider the signal model given in Equation (1), with $k_0 = 0$ and $a(n)$ a zero-mean Gaussian stationary process with correlation $c_a(\tau)$ and pseudo-correlation $p_a(\tau)$. For the sake of simplicity, we further assume that the noise statistics, i.e., $\{c_a(\tau), p_a(\tau)\}_{\tau\in\mathbb{Z}}$ and σ_w^2, are known at the receiver. This assumption was made in [49], and was partially made in [28], for deriving Barankin bounds (BB) because the computational and analytical complexities are otherwise too high. It can also be noted that the Cramer-Rao bound for frequency estimation is insensitive to the knowledge of the noise statistics as soon as the number of samples is large enough (see [31] and the Gaussian-Cramer-Rao-bound discussion above). We thus can expect that the error induced by neglecting the estimation step with respect to the noise statistics will be sufficiently small so that our further conclusions still hold in the case of unknown noise statistics.

To well understand the interest of bounds other than the Cramer-Rao bound, let us consider the following

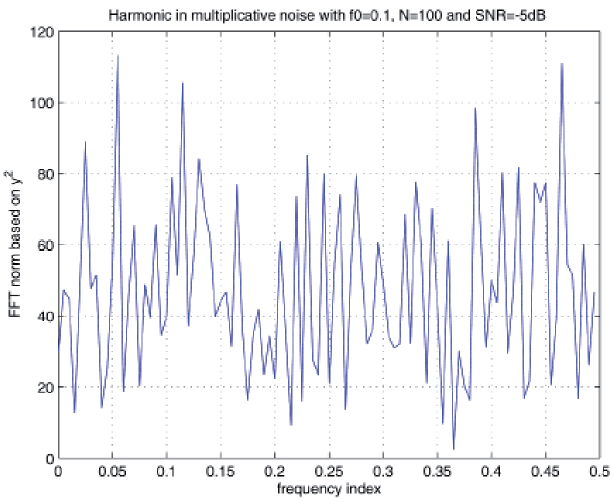

Figure 2a. The cost function $F(\varphi)$ as a function of φ for SNR = −5 dB.

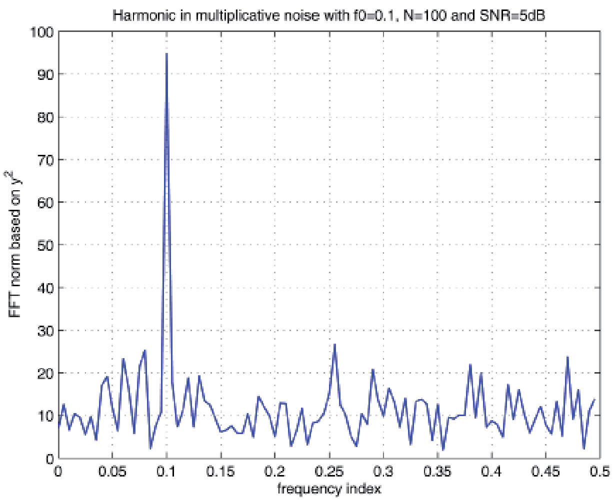

Figure 2b. The cost function $F(\varphi)$ as a function of φ for SNR = 5 dB.

example. The signal model is the model from Equation (1) with $a(n)$ a real-valued Gaussian process. To estimate the frequency, as $a(n)$ is noncircular (since real-valued), one can use the so-called square-power estimator [47], defined as follows:

$$\hat{\varphi}_1 = \arg\max_{\varphi} \underbrace{\left| \frac{1}{N} \sum_{n=0}^{N-1} r(n)^2 e^{-2j\pi(2\varphi)n} \right|^2}_{F(\varphi)} .$$

In Figure 1, we plot the mean-square estimation error of this estimator, and the modified and Gaussian Cramer-Rao bound, as functions of the SNR, when $N = 100$. We observed that at high SNR, the estimator was powerful and even efficient (the mean-square estimation error equals the Cramer-Rao bound). In contrast, at low SNR, there was a large mismatch between the mean-square estimation error and the Cramer-Rao bound. The question is, is the considered estimator not relevant at low SNR, or is the Cramer-Rao bound not tight enough at low SNR? We will show that the Cramer-Rao bound is not tight enough. To demonstrate that, we introduce other lower bounds on the mean-square estimation error that are much tighter at low SNR than the Cramer-Rao bound.

We can now attempt to understand why the Cramer-Rao bound is not tight enough at low SNR. In Figure 2, we plot the cost function, $F(\cdot)$, of the square-power estimator for $\mathrm{SNR} = -5$ dB (Figure 2a) and for $\mathrm{SNR} = 5$ dB (Figure 2b). The frequency sought is $\varphi_1 = 0.1$. We remark that at high SNR, the peak around the true value of the frequency was well detected, whereas at low SNR, there was a mis-detection of the peak, which significantly degraded the performance. Consequently, the performance degradation was due to a higher peak far away from the true frequency. These "bad" realizations are called "outliers." By inspecting in detail the Fisher information matrix from Equation (3), we remark that it depends on the behavior of the likelihood function around the true frequency, since the derivative functions involved are calculated at the true frequency. The Cramer-Rao bound is therefore unable to take into account the mis-detection of the peak, and automatically assumes a correct detection of the peak, even if it is wrong. At low SNR (when the mis-detection of the peak occurs), the Cramer-Rao bound is thus truly too optimistic.

We are now interested in another bound that inspects the likelihood function around the true frequency, but not only there. We therefore introduce the following set of so-called "test-points,"

$$\left\{ \phi(k) = \left[\phi_0(k), \phi_1(k) \right]^{\mathrm{T}} \right\}_{1 \le k \le n} ,$$

at which the likelihood function will be evaluated. We are now able to define the Barankin bound of order p as follows:

$$\mathrm{BB}_p(\varphi_0, \varphi_1) = \sup_{\mathcal{E}} S_p(\mathcal{E}) ,$$

where

$$S_p(\mathcal{E}) = \mathcal{E}\left[\mathbf{B}(\mathcal{E}) - \mathbf{1}_p \mathbf{1}_p^{\mathrm{T}} \right]^{-1} \mathcal{E}^{\mathrm{T}} ,$$

with

$$\mathcal{E} = \left[\phi(1) - \mathbf{u}^{\mathrm{T}}, \ldots, \phi(p) - \mathbf{u}^{\mathrm{T}} \right]$$

and

$$\mathbf{1}_p = \mathrm{ones}(p,1) .$$

The term $\sup$ stands for the smallest upper bound on the set $\mathcal{E}$. Furthermore, $\mathbf{B}(\mathcal{E}) = \left(B_{k,l} \right)_{1 \le k,l \le p}$ is the following $p \times p$ matrix:

$$B_{k,l} = \mathbb{E}\left\{ L\left[\mathbf{r}, \mathbf{u}, \phi(k) \right] L\left[\mathbf{r}, \mathbf{u}, \phi(l) \right] \right\} ,$$

with

$$L\left[\mathbf{r}, \mathbf{u}, \phi(k) \right] = \frac{p\left[\mathbf{r} \mid \phi(k) \right]}{p(\mathbf{r} \mid \mathbf{u})} .$$

The mean-square estimation error of any unbiased estimator is greater than the Barankin bound of any order p ([40]). From an asymptotic point of view (as $p \to \infty$), the Barankin bound is even the tightest lower bound that one can find [27, 28]. As for the choice of the test points, it is usual to consider the following structure for $\mathcal{E}$ [28, 29]:

$$\mathcal{E} = \begin{bmatrix} \phi_0 - \varphi_0 & 0 \\ 0 & \phi_1 - \varphi_1 \end{bmatrix} = \mathrm{diag}(\varepsilon_0, \varepsilon_1) . \qquad (17)$$

Our main concern hereafter is to derive a closed-form expression for the matrix $\mathbf{B}$ for such test points.

Let us now remind the reader of some notation. The covariance matrix $\tilde{\mathbf{C}}_{\mathbf{r}}(\phi)$ of the multivariate process $\tilde{\mathbf{r}}$ can be written as follows:

$$\tilde{\mathbf{C}}_{\mathbf{r}}(\phi) = \tilde{\mathbf{S}}(\phi)\left(\tilde{\mathbf{C}}_{\mathbf{a}} + \sigma_w^2 \mathbf{I}_{2N} \right) \tilde{\mathbf{S}}^{\mathrm{H}}(\phi) , \qquad (18)$$

where

$$\tilde{\mathbf{S}}(\phi) = \begin{bmatrix} \mathbf{S}(\phi) & \mathbf{0}_N \\ \mathbf{0}_N & \bar{\mathbf{S}}(\phi) \end{bmatrix} .$$

After straightforward algebraic manipulations, we finally obtain

$$B_{k,l}=\begin{cases}\dfrac{1}{\sqrt{\det\left(\mathbf{Q}_{k,l}\right)}} & \text{if } \mathbf{Q}_{k,l}>0\\ +\infty & \text{otherwise}\end{cases},$$

with

$$\mathbf{Q}_{k,l}=\left\{\tilde{\mathbf{C}}_{\mathbf{r}}\left[\phi(k)\right]^{-1}+\tilde{\mathbf{C}}_{\mathbf{r}}\left[\phi(l)\right]^{-1}\right\}\tilde{\mathbf{C}}_{\mathbf{r}}(\mathbf{u})-\mathbf{I}_{2N}.$$

These expressions were introduced in [49] and [28] in a slightly different form (due to the circularity assumption on the non-constant amplitude), and in [31] for the general case.

We now just focus on our main parameter of interest: the frequency, φ_1. For the standard test points described in Equation (17), the Barankin bound for φ_1 takes the following form [28]:

$$\mathrm{BB}(\varphi_1)=\sup_{\varepsilon_0,\varepsilon_1}\frac{\varepsilon_1^2}{\left(B_{1,1}-1\right)-\dfrac{\left(B_{0,1}-1\right)^2}{B_{0,0}-1}}.$$

The term $\left(B_{0,1}-1\right)^2/\left(B_{0,0}-1\right)$ represents the loss in performance due to joint phase and frequency-parameter estimation.

We remark that strictly speaking, the Barankin bound is not obtained in closed form, since the maximum operator still occurs. Nevertheless, the existing expressions enable us to very quickly compute the Barankin bound.

As we will see in the simulation section, the Barankin bound enables us to partially predict the outliers effect, i.e., the mismatch between the Cramer-Rao bound and the real estimator's performance. Consequently the poor estimation performance of the standard square-power estimator (well adapted to BPSK or a real-valued Gaussian process) is shown to be connected to the poor tightness of the Cramer-Rao bound. Decreasing the gap between the Cramer-Rao bound and the estimator performance for a given number of samples at very low SNR is thus impossible. The Cramer-Rao bound is too optimistic in such a context, and has to be replaced with the Barankin bound.

6. Deriving the Ziv-Zakai Bound

To analyze the mismatch between the Cramer-Rao bound and the real estimator's performance, we have considered the so-called Barankin bound in the previous section. Even though this Barankin bound is much tighter than the Cramer-Rao bound and roughly predicts the outliers effect, there is still a mismatch between bound and estimator performance. In this section, we will therefore introduce a third, much-more-powerful bound, the Ziv-Zakai bound (ZZB).

We note that the Ziv-Zakai bound needs a new paradigm on the parameters sought: the Bayesian approach. Unlike what was previously done, we have to consider the parameters sought as a realization of a random variable. This random variable is further described by a distribution, which characterizes the *a priori* information available on the parameters sought. For instance, for frequency-offset estimation, we only know that the frequency is normalized, and thus it may uniformly take values in the interval $\left[-1/2,1/2\right]$.

In [20, 50], it was proven that the following inequality holds for any vector $\mathbf{z}=\left[z_0,z_1\right]$:

$$\mathbf{z}\mathbf{E}_{\mathbf{u}}\mathbf{z}^{\mathrm{T}}\geq\int_0^{\infty}\Delta\left[\max_{\substack{(\varepsilon_0,\varepsilon_1)\\ z_0\varepsilon_0+z_1\varepsilon_1=\Delta}} f\left(\varepsilon_0,\varepsilon_1\right)\right]d\Delta, \quad (19)$$

where $\mathbf{E}_{\mathbf{u}}$ denotes the error-correlation matrix related to the estimation of a random variable $\mathbf{u}=\left[\varphi_0,\varphi_1\right]$, and

$$f\left(\varepsilon_0,\varepsilon_1\right)=\int\min\left[p(\mathbf{u}),p(\mathbf{u}+\varepsilon)\right]P_e\left(\mathbf{u},\mathbf{u}+\varepsilon\right)d\mathbf{u}, \quad (20)$$

with $\varepsilon=\left[\varepsilon_0,\varepsilon_1\right]$.

The function $p(\mathbf{u})$ is the a priori density function of the bivariate parameter $\mathbf{u}$, and $P_e\left(\mathbf{u},\mathbf{u}+\varepsilon\right)$ is the error probability when the optimal detector (namely, the ML detector) is used to decide between the following two hypotheses:

$$\begin{cases}H_0: y(n)=a(n)e^{2i\pi(\varphi_0+\varphi_1 n)}+w(n)\\ H_1: y(n)=a(n)e^{2i\pi\left[(\varphi_0+\varepsilon_0)+(\varphi_1+\varepsilon_1)n\right]}+w(n)\end{cases}$$

where hypotheses H_0 and H_1 are equally likely.

The right-hand side of Equation (19) is called the Ziv-Zakai bound. By inspecting Equation (19), one can remark that the likelihood of $\mathbf{u}$ is scanned over the entire search interval of $\mathbf{u}$, as is also the case for the Barankin bound [31]. Once again, this contrasts with the Cramer-Rao bound, where the likelihood function is only evaluated around the true point. We therefore expect that the Ziv-Zakai bound can predict the outliers effect at low SNR.

Let us focus now on the Ziv-Zakai bound for φ_1, which is obtained by setting $\mathbf{z}=\left[0,1\right]$. Therefore,

$$\mathrm{ZZB}(\varphi_1) = \int_0^{\infty} \varepsilon_1 \left[\max_{\varepsilon_0} f(\varepsilon_0, \varepsilon_1) \right] d\varepsilon_1 .$$

Actually, the mean-square estimation error of any (even biased) estimator for the frequency is greater than the $\mathrm{ZZB}(\varphi_1)$ [20].

The key task now is to express the function $f(\cdot)$ in closed form. After some simple derivations, one can see that $P_e(\mathbf{u}, \mathbf{u}+\varepsilon)$ is independent of $\mathbf{u}$, so that it can be denoted by $P_e(\varepsilon_0, \varepsilon_1)$. As a consequence, we have [50]

$$f(\varepsilon_0, \varepsilon_1) = g(\varepsilon_0, \varepsilon_1) P_e(\varepsilon_0, \varepsilon_1),$$

where

$$g(\varepsilon_0, \varepsilon_1) = \int \min\left[p(\mathbf{u}), p(\mathbf{u}+\varepsilon) \right] d\mathbf{u} .$$

Since we have no a priori information on $\mathbf{u}$, we assume that φ_0 and φ_1 are uniformly distributed over $[0, 1/2]$, i.e., the a priori distribution of the parameters of interest $p(\mathbf{u})$ is flat. We consider the interval $[0, 1/2]$ rather than $[-1/2, 1/2]$ because the phase and the frequency can only be estimated modulo $1/2$ when multiplicative noise occurs [50]. Consequently,

$$g(\varepsilon_0, \varepsilon_1) = (1/2 - \varepsilon_0)(1/2 - \varepsilon_1).$$

This leads to

$$\mathrm{ZZB}(\varphi_1)$$

$$= \int_0^{1/2} (1/2 - \varepsilon_1) \varepsilon_1 \max_{\varepsilon_0} \left[(1/2 - \varepsilon_0) P_e(\varepsilon_0, \varepsilon_1) \right] d\varepsilon_1 .$$

The rest of the section deals with the evaluation of $P_e(\varepsilon_0, \varepsilon_1)$. After tedious algebraic derivations that can be found in [53], we have

$$P_e(\varepsilon_0, \varepsilon_1) = \mathrm{Prob}\left[\sum_{n=m}^{2N-1} \lambda_n^{(+)} v_n^2 < \sum_{n=0}^{m-1} \lambda_n^{(-)} v_n^2 \right], \quad (21)$$

where

- $\{v_n\}$ is a real-valued i.i.d. (independent identically distributed) Gaussian random process with zero mean and unit variance.
- The $-\lambda_n^{(-)}$ for $n = 0, \cdots, m-1$ (with $\lambda_n^{(-)} > 0$) are the m negative eigenvalues, and $\lambda_n^{(+)} \geq 0$ for $n = m, \cdots, 2N-1$ are the positive or null eigenvalues of the $2N \times 2N$ matrix $\mathbf{T}(\varepsilon)$, defined as follows:

$$\mathbf{T}(\varepsilon) = \left[\breve{\mathbf{C}}_{\mathbf{r}}(\varepsilon)^{-1} - \breve{\mathbf{C}}_{\mathbf{r}}(0)^{-1} \right] \breve{\mathbf{C}}_{\mathbf{r}}(0),$$

where $\breve{\mathbf{C}}_{\mathbf{r}}(\varepsilon) = \mathbb{E}\left[\breve{\mathbf{r}}^{\mathrm{T}} \breve{\mathbf{r}} \right]$ with $\breve{\mathbf{r}} = \{\Re[\mathbf{r}], \Im[\mathbf{r}]\}$ and $\mathbf{r}$ is the received signal disturbed by phase ε_0 and frequency ε_1.

We now wish to derive a closed-form expression for the following term:

$$P_e(\varepsilon_0, \varepsilon_1) = \mathrm{Prob}(p_+ < p_-), \quad (22)$$

where $p_\pm = \sum \lambda_n^{(\pm)} v_n^2$ is a weighted sum of squared independent Gaussian variables. Notice that by construction, p_+ and p_- are independent.

If $\lambda_n^{(+)} = \lambda^{(+)}$ (respectively, $\lambda_n^{(-)} = \lambda^{(-)}$) for all corresponding n, then p_+ (respectively, p_-) obeys a χ^2 distribution with $(2N - m)$ (respectively, m) degrees of freedom. However, if the weighting coefficients are different, the $p_\pm$ s are not χ^2 distributed anymore. Further, expressing the distribution of $p_\pm$ in closed form is not tractable. Nevertheless, it can be well approximated by means of the Gamma distribution [51]. We recall that the Gamma distribution, denoted $\mathcal{G}(\aleph, \wp)$, is defined as follows:

$$P_{\aleph,\wp}(x) = \frac{x^{\aleph-1}}{\Gamma(\aleph)\wp^{\aleph}} e^{-x/\wp},$$

where $\Gamma(\cdot)$ is the Gamma function.

Hence, the distribution of $p_\pm$ is next approximated by the Gamma distribution, the first and second moments of which are equal to those of $p_\pm$. We thus obtain

$$p_+ \sim \mathcal{G}(\aleph_+, \wp_+)$$

and

$$p_- \sim \mathcal{G}(\aleph_-, \wp_-),$$

with

$$\aleph_+ = \frac{1}{2} \frac{\left[\sum_{n=m}^{2N-1} \lambda_n^{(+)} \right]^2}{\sum_{n=m}^{2N-1} \lambda_n^{(+)2}}$$

and

$$\wp_+ = 2 \frac{\sum_{n=m}^{2N-1} \lambda_n^{(+)2}}{\sum_{n=m}^{2N-1} \lambda_n^{(+)}},$$

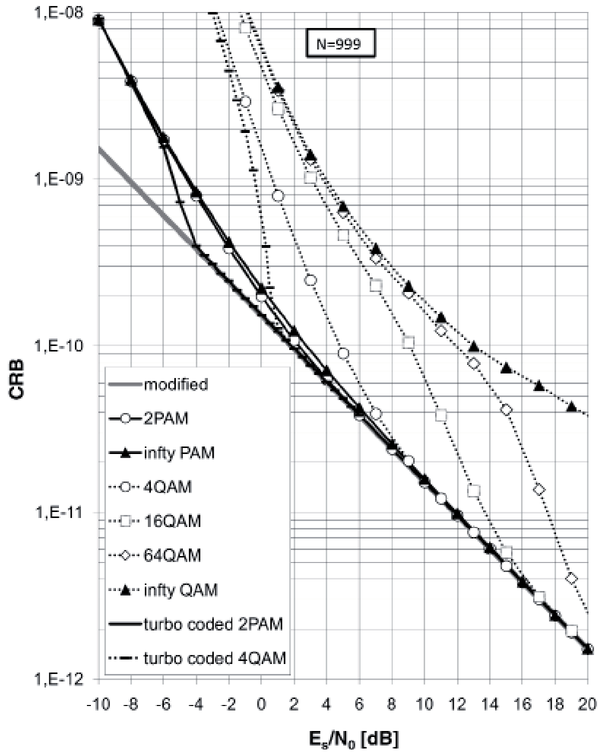

Figure 3. The Cramer-Rao bounds related to the estimation of φ_1 resulting from the observation model, Equation (1), as a function of the SNR, E_s/N_0, for random, linear, MPAM, and MQAM, coded and un-coded, modulations.

and

$$\aleph_- = \frac{1}{2}\frac{\left[\sum_{n=0}^{m-1}\lambda_n^{(-)}\right]^2}{\sum_{n=0}^{m-1}\lambda_n^{(-)2}}$$

and

$$\wp_- = 2\frac{\sum_{n=0}^{m-1}\lambda_n^{(-)2}}{\sum_{n=0}^{m-1}\lambda_n^{(-)}}.$$

As $p_\pm$ is now assumed Gamma distributed, Equation (22) can be simplified. Indeed, by using the fact that the square root of a Gamma-distributed random variable is Nakagami distributed, and by using Equation (46) in [52], we have that

$$P_e(\varepsilon_0,\varepsilon_1) = \left(\frac{\wp_+}{\wp_-}\right)^{\aleph_+}\frac{\Gamma(\aleph_+ + \aleph_-)}{\aleph_+\Gamma(\aleph_+)}$$

$${}_2F_1\left(\aleph_+ + \aleph_-, \aleph_+, \aleph_+ + 1; -\frac{\wp_+}{\wp_-}\right),$$

where ${}_2F_1(\cdot)$ is the hypergeometric function.

The above expression for $P_e(\varepsilon_0,\varepsilon_1)$ represents the main available result on the Ziv-Zakai bound derivations [53]. Although this expression is not interpretable, its numerical computation will provide interesting results, as seen below. Notice also that the expressions obtained for the Ziv-Zakai bound are not anymore a bound, strictly speaking, since we are not able to prove that the approximate expressions are less than the exact (but unavailable) bound. However, by checking the approximation numerically, we have observed that the approximation is very tight.

7. Simulation Results

7.1 Non-Constant Complex Amplitude = Digital Data Symbol

Figure 3 presents some numerical results for the true Cramer-Rao bound regarding the estimation of the frequency offset from the observation of $N = 999$ linearly modulated signal samples that were obtained by means of computer simulations. The following signaling constellations, Ω, were considered:

- *M*-ary *pulse-amplitude modulation*} (*M*-PAM) for which

$$\Omega = \sqrt{\frac{3}{\left(M^2-1\right)}}\mathcal{I}_M.$$

- *M*-ary *quadrature-amplitude modulation* (*M*-QAM) for which

$$\Omega = \left\{\omega : \Re\{\omega\}, \Im\{\omega\} \in \sqrt{\frac{3}{2(M-1)}}\mathcal{I}_{\sqrt{M}}\right\},$$

where $\Re\{\cdot\}$ and $\Im\{\cdot\}$ denote the real and the imaginary parts of a complex number.

In the above,

$$\mathcal{I}_m = \{\pm 1, \pm 3, \ldots, \pm(m-1)\}. \qquad (23)$$

We further considered uncoded and turbo-coded linear modulation. The turbo-coded transmission scheme encompasses the parallel concatenation of two identical binary 16-state rate-1/2 recursive systematic convolutional encoders, with generator polynomials $(21)_8$ and $(37)_8$ in octal notation, via a pseudo-random interleaver with block length N_b information bits. An appropriate puncturing pattern, so that the block at the turbo-encoder output comprises N_c coded bits, was used. This binary turbo code was followed by conventional Gray-mapped 2PAM or

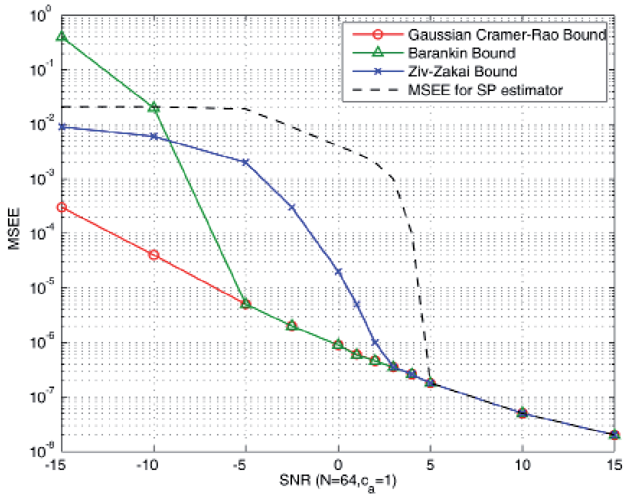

Figure 4. The mean-square estimation error as a function of the SNR.

4QAM modulation, giving rise to a block of N random data symbols, with $N = N_c = 3N_b$ for the case 2PAM, and $N = N_c/2 = N_b$ for the case of 4QAM.

Our simulation results confirmed that the high-SNR limit of the Cramer-Rao bounds equals the modified Cramer-Rao bound. Comparing the Cramer-Rao bounds for coded and uncoded transmission, we observed that for a given constellation type, they were equal at sufficiently high SNRs. However, at lower SNRs there was a gap between the Cramer-Rao bounds for coded and uncoded transmissions. When E_s/N_0 decreased, a point $(E_s/N_0)_{thr}$ was reached where the Cramer-Rao bounds started to diverge from their high-SNR limit. For coded transmission, $(E_s/N_0)_{thr}$ corresponded to a coded BER (bit-error rate) of about 10^{-3}. For uncoded transmission, $(E_s/N_0)_{thr}$ corresponded to an uncoded BER of about 10^{-3}, and consequently exceeded $(E_s/N_0)_{thr}$ for coded transmission by an amount equal to the coding gain.

For uncoded transmission, the following observations can further be made:

- For both constellation types (PAM, QAM), we observed that for a given value of E_s/N_0, the Cramer-Rao bound increased with M, which indicated that for the larger constellations, carrier recovery was inherently harder to accomplish. This effect was clearly evident for MQAM, in which case the curves corresponding to large M exhibited an almost horizontal portion, but was almost unnoticeable for MPAM. Figure 3 also shows the limiting curve for M approaching infinity. This situation corresponded to data symbols that were continuous random variables, that were uniformly distributed in the interval $\left[-\sqrt{3}, \sqrt{3}\right]$ for PAM, and in a square with side $\sqrt{6}$ for QAM. In the case of infinite-size constellations, the Cramer-Rao bounds do not necessarily converge to the corresponding modified Cramer-Rao bounds for large SNR, according to [11]. This is due to the non-diagonal nature of the Fisher information matrix, related to the joint likelihood function $p(\mathbf{r}|\mathbf{a},\mathbf{u})$ of $\mathbf{a}$ and $\mathbf{u}$, with $\mathbf{u} = [\varphi_0, \varphi_1]$.

- For finite M, the Cramer-Rao bound does converge to the modified Cramer-Rao bound when E_s/N_0 is

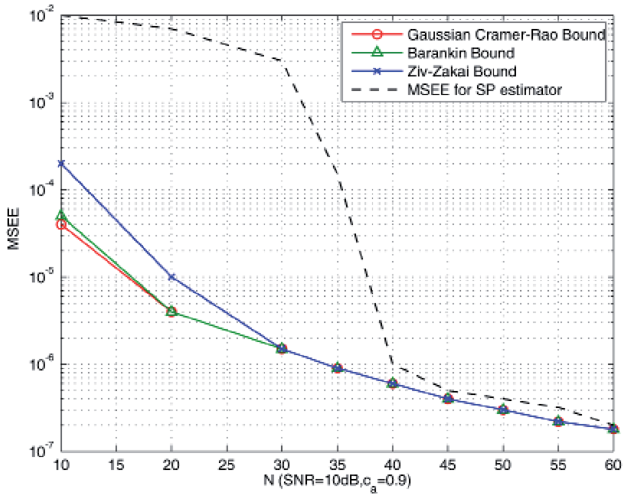

Figure 5. The mean-square estimation error as a function of N.

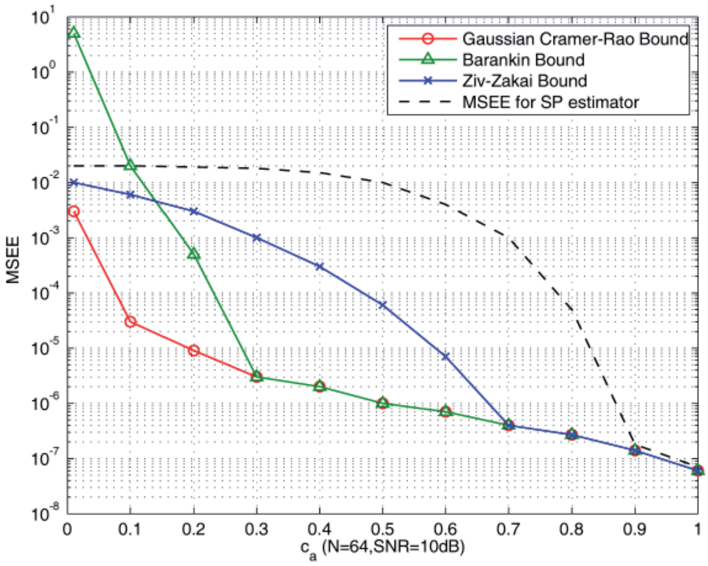

Figure 6. The mean-square estimation error as a function of c_a.

sufficiently large. The value of E_s/N_0 at which Cramer-Rao bound is close to the modified Cramer-Rao bound increases by about 6 dB when M doubles (PAM) or quadruples (QAM). This indicates that for uncoded pilot-symbol-free transmission, the convergence of the Cramer-Rao bound to the modified Cramer-Rao bound is mainly determined by the value of $\frac{E_s}{N_0}(d_M)^2$, with d_M denoting the minimum Euclidean distance between the constellation points. Furthermore, at the normal operating SNR of uncoded digital-communication systems, the Cramer-Rao bounds turn out to be very well approximated by the corresponding modified Cramer-Rao bounds.

7.2 Non-Constant Complex Amplitude = Gaussian

The multiplicative noise, $a(n)$, is hereafter assumed to be a white non-circular Gaussian process with zero mean, unit variance, and pseudo-variance $c_a = \mathbb{E}\left[a(n)^2\right]$. For the sake of simplicity, we also assume that the real part of $a(n)$ is independent of its imaginary part. This implies that c_a is real-valued. If $c_a = 0$, then $a(n)$ is circular; if $c_a = 1$, then $a(n)$ is real-valued. Thus, c_a quantifies the non-circularity rate of $a(n)$. We also set $\text{SNR [dB]} = 10\log_{10}\left(1/\sigma_w^2\right)$.

In each figure, we display four curves. Dashed lines correspond to the empirical mean-square estimation error for the well-known square-power (SP) estimate [1, 31, 34]. Solid lines with star-shape markers represent the Ziv-Zakai bound. Solid lines with triangular-shaped markers represent the Barankin bound. Solid lines with circular-shaped markers represent the Cramer-Rao bound [31, 53].

In Figure 4, we plot all the curves as a function of the SNR with $N = 64$, $c_a = 1$. We observed that the well-known outliers effect occurred at low and medium SNR [33]. We also observed that the Ziv-Zakai bound is significantly tighter than the Barankin bound. The SNR threshold corresponding to the SP-based estimate was much larger than that observed with the Barankin bound, while the threshold value predicted by the Ziv-Zakai bound was quite close to that obtained empirically with the square-power estimate. As a consequence, the Ziv-Zakai bound seems to be more powerful than the Barankin bound.

In Figure 5, we plot the curves as a function of N with SNR = 10 dB, $c_a = 0.9$. Even though the Ziv-Zakai bound offered a more realistic value for the N threshold than the Barankin bound, the mismatch between the Ziv-Zakai bound and the square-power mean-square estimation error performance was still quite large.

In Figure 6, the curves are displayed as a function of c_a with $N = 64$, SNR = 10 dB. One could notice that the more $a(n)$ was non-circular (i.e., c_a increased), the better the estimation performance. Furthermore, the outliers effect rapidly degraded the performance if $a(n)$ was not sufficiently non-circular. The figure confirmed that accurate frequency estimation is really difficult to achieve when the white signal is not sufficiently non-circular.

8. Conclusions

In this tutorial, we have focused on the derivation and the analysis of fundamental lower bounds on the achievable mean-square estimation error for estimating the frequency and the phase of a received signal, when the complex amplitude of the signal is non-constant and unknown. In particular, the following application fields have been considered: digital communications, direction-of-arrival estimation, and Doppler radar. An overview of lower bounds (the Cramer-Rao bound, modified Cramer-Rao bound, Barankin bound, and Ziv-Zakai bound), with their respective interests and their associated derivations in closed form for various cases, has been presented.

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The Role of Radio Science in Disaster Management

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D.G. Cole

Abstract

Recent major natural disasters have brought a greater awareness of the social and economic deprivations that follow global emergencies. Among other organizations, URSI (International Union of Radio Science) is assessing its contribution, past and future, to the mitigation of natural and human-induced hazards.

This paper summarizes the way in which radio science has contributed to lessening the impact of disasters, both in the past and future. Its objective is to encourage all radio scientists to think how they can contribute further to alleviate the impact of disasters.

1. Introduction to URSI Inter-Commission WG1

Between 1975 and 2009, over 10,000 natural disasters worldwide killed more than 2,500,000 people, and produced estimated damages of over 1.7 trillion US dollars. Earthquakes, landslides, tropical cyclones, severe storms, floods, and infectious diseases were the major causes [1].

Science and technology working with society can reduce the risk and impact of disasters.

Radio science pervades society, and has an integral role in disaster management and mitigation that is often taken for granted. Radio is a vital element in monitoring the environment, and in feeding data to prediction models that are a major factor in safety and economic wellbeing [2]. When the telecommunication infrastructure is significantly or completely destroyed in a disaster, then radio communications (especially radio-amateur and satellite services) become important for disaster-relief operation. Building on radio research – which the URSI Commissions embrace – the International Telecommunication Union (ITU) and World Meteorology Organization (WMO) have developed recommendations, reports, and handbooks related to the use of radio communications to reduce the impact of natural and manmade disasters (including the anticipated effects of climate change) [3, 4].

In response to an increasing vulnerability to global disasters, the 2008 URSI General Assembly created a Working Group on Natural and Human Induced Hazards (WG1) [5], with the following terms of reference:

a) To study, within the URSI area of competence, methods and strategies related to natural and human-induced environmental hazards and disasters, such as:

   (i) Communication systems suitable for fast-response disaster relief;

   (ii) The development and application of remote-sensing products and other global data for monitoring and alerting;

   (iii) The evaluation of long-term and short-term risks of disasters, and;

   (iv) The description of the environment disturbances resulting from disasters;

b) To provide support to initiatives taken in the area of risk management and relief related to natural and human-induced catastrophes and disasters, particularly by developing countries.

This paper provides some background and ideas on ways in which URSI has contributed, and will continue to contribute, to the reduction of public risk from natural and human-induced hazards. The ideas and techniques presented here will be further refined. Readers are encouraged to contact the WG1 Chair and lead author (Phil Wilkinson) with any comments, ideas, and additional references, including Web references. This information will be added to the Working Group Web site, shortly to be linked to the URSI Web site (http://www.ursi.com).

2. Stages in Hazard Risk Reduction

It is practically impossible to prevent the occurrence of natural disasters and subsequent damage. However, it is possible to reduce the impact of disasters by the following:

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a) Prediction and prevention of a disaster situation

b) Reducing local disaster risks

c) Immediate disaster response

d) Disaster relief support

e) Learning and disseminating information from a disaster

Radio science plays a part in all these stages.

2.1 Prediction and Prevention of a Disaster Situation

Sound preparation prior to a disaster is mandatory. The Internet is a valuable communication resource for providing this information to both disaster specialists and the general public. Prediction models for many natural disasters can give valuable advance warning, particularly where meteorological and hydrological factors are involved. Web sites providing geospatial and remote-sensing data and services (e.g., see [6, 7]) can provide regional portals for current warnings of predicted potential and measured situations around the world for all types of hazards, such as floods, windstorms, firestorms, earthquakes, volcanic eruption, and tsunamis. Ideally, these Web sites will be simple to use, low impact, and available where risks are high.

Obviously, prediction models depend on current data, rapid analysis, and adequate communications. Predictions will be integrated with communications and broadcasting services to facilitate the production of alerts. Many hazards are related to extreme weather conditions, and therefore national weather agencies have a major role to play in coordinating data and disseminating alerts (e.g., [8-10]). Fast, accurate, and automatic data communication is essential in monitoring the development of a disaster situation. Web sites need to be up-to-date, and to provide information both for those likely to be directly affected, and those who may need to be informed about the developing situation. In addition, advice for personal preparations and hazard prevention can be provided for emerging risks.

URSI is in a position to provide specialist information on the background to radio-related information and prediction models appropriate to each type of hazard.

2.2 Reducing Local Disaster Risks

Risks need to be predicted and reduced through planning, preventative strategies, and by constantly assessing local environmental conditions. Improving realistic risk prediction is a major challenge in the mitigation of natural hazards. In general, the risk will probably need to be high before any expensive precautions are taken.

Disaster risk assessment will use global regional and local environmental models, together with current data. Construction of data ensembles will use a wide range of radio techniques for both their collection and transmission. Robust, high-volume, fast, and secure networks provide data for analysis. In addition, the data will need to be subject to quality control and interpretation, both areas requiring professional skills and research in areas of increasing complexity. The outcome of this work will need to be passed back to the people concerned.

URSI Commissions have a role here, also. For instance, URSI Commission F provides image collection, processing, quality control, and interpretation skills, and other Commissions provide knowledge in the areas of communications.

2.3 Immediate Disaster Response

When an event occurs that may cause a disaster, urgency enters the scene. Assessment and evolving strategies depend on instantaneous, accurate information. Disaster management calls for an immediate assessment of the scale of the disaster, the likely losses and the potential danger to rescue teams. Advances in remote sensing and satellite communication are improving the speed and accuracy of this information gathering, making effective loss estimates immediately available after a disaster possible, enhancing the effectiveness of mitigation.

As soon as possible, a control and command center is established near the disaster zone. This is vital for efficient management. The control center will bring data and communications into and out of the region, provide communications with transport and personnel, as well as provide robust communications back to the national base. This is a well-established radio-science role.

Take communications as an example. An earthquake strikes a township of several tens of thousands. The development of communications over the period of the disaster is likely to be as follows:

- Immediate concern that a disaster may have happened and attempts to communicate via a public network *(immediate use of any remaining telephone, mobile, and HF communication infrastructure)*
- Establishment of HF communication links if such exist within the disaster community *(immediate timeframe – often radio amateurs – to be formalized later)*

- Emergency frequency-coordination plan established to allow Earth-space, ground communications, and aviation services to operate *(first day)*

- VHF and solar-powered satellite emergency communication systems are dropped into the zone by air to establish a general communications HQ *(first few days)*

- Other sensors and equipment are dropped into the zone with the need for communications network throughout the disaster zone and back to the zone communications HQ *(first week)*

- Data and information communication increase between zone HQ and outside assistance *(first week)*

- Transport (air, sea, and land) control and management now requires greater bandwidth and volume of radio traffic

- Media coverage from the zone develops and requires coordination with other frequencies in operation

- Internal communication system is re-established so that communication within the community and outside is enabled.

URSI has already played an important role by providing a platform for the radio research that lead to the ITU-R procedures for radio-frequency allocations used in emergency situations.

Natural fires are another threat in many countries. In Australia, the summer bushfire season mobilizes a range of rescue services, and local radio broadcasts alert the public when weather conditions escalate risks. A bushfire can cover wide regions of land and, when fanned by gusty winds, can travel at impressive speeds, presenting immense danger and significant communication and logistical problems. During moderate fires, houses are saved sometimes with a meter or two to spare, and fires can be guided and contained. Yet, if fires escalate to catastrophic levels, as they did in Victoria in February 2009, then property losses are high (2000 residences) and lives are lost (over 200 people died). Fires of catastrophic magnitude create their own propagation conditions, destroy infrastructure causing congestion and failure of the remaining facilities, and present major communication and logistical problems. In the case of the Victoria fires, HF radio proved valuable in difficult terrain [11]. Unlike earthquakes, fire disasters last many days, and the intensity of the fire depends on the weather. As the fire intensity increases, our radio-science knowledge, embedded in communication networks and links, is stressed: an obvious area for further research.

There is still much to learn and apply in coping with radio performance in extreme conditions. This is especially true of the high-stress periods during the urgent first stages of a developing disaster, when robust, highly dependable stable techniques are essential.

2.4 Disaster Relief Support

Disaster relief requires realistic assessment of priorities both inside the disaster zone and outside, within the national/international community. Relief follows the containment of the hazard. Adequate relief depends upon reliable decision making, resources, equipment, and transport and information flow.

By this stage, radio communications will have been stabilized, and relief is likely to focus on the search for survivors, to provide medical treatment, and emergency housing and hygiene. In a modern disaster environment, this is likely to include radio and radar-related technical support, sophisticated techniques such as telemedicine, and high-definition imagery and automated systems warning of further hazards.

During this phase of a disaster, when control is more stable, it may be efficient to employ more specialized techniques, possibly testing some of the latest technology in highly stressful conditions.

2.5 Learning and Disseminating Information from a Disaster

Every hazardous situation provides lessons for the future. These must be captured and disseminated in a way that can be understood, not only by those directly affected, but also by those outside the immediate disaster region, and especially those in other potential threat zones. Appropriate authorities need to respond to the lessons emanating from each event with decisions.

URSI has many of the skills required to advise authorities on matters such as improved radio communication and warning systems. Furthermore, through its direct links to the academic world, URSI is a natural conduit to spread educational messages.

In each of the above stages of a disaster, reliable communications with adequate information and data flow into and out of the disaster zone are vital. Rapid response can save lives.

3. Possible Radio-Science Contributions in Disaster Management

URSI Working Group 1 on Natural and Human Induced Hazards was set up in response to the ICSU research policy of characterizing hazards, vulnerability, and risk, understanding decision-making in risk management, and reducing risk through knowledge-based actions [12]. The knowledge in URSI Commissions has provided and will continue to provide assistance in reducing the impact of natural and manmade hazards.

We need an audit of the radio-science techniques that are in use, or could be used, during disasters. Are these techniques used effectively? Could a small amount of advice make a significant difference? Are the special conditions of a disaster matched by equipment performance? For instance, are propagation conditions the same? Has the noise environment changed? How do damaged networks respond to congestion? Will imagery provide effective disaster-management advice? There are many questions like these that need answers. On the one hand, a better understanding of disaster environments may stimulate thoughts about better operational practices. On the other hand, better practices may evolve from the application of new radio-science research.

Radio science can contribute to disaster relief across a spectrum of inputs. Table 1 provides a rudimentary summary of these inputs. Running horizontally, the movement flows from well-known, standard techniques, through the less conventional, more-modern applications, through to the domain of pure research. Vertically, the table suggests how the different domains interact with a disaster scenario, and how they may contribute better in the future.

Bearing these ideas in mind, some of the URSI research studies that have contributed to the task of hazard reduction, or might contribute, are listed below. They consist of research in progress, studies completed in recent years, and work that should be considered in the future. The list is certainly not exhaustive, and readers are encouraged to share their creativity in this area.

3.1 Radio Communication Systems

Radio communication systems have advanced in innumerable ways over the last decade, both in techniques and also in the uptake of new technology. Below are a few items that might contribute to improving communications in a disaster situation.

- High-gain antennas: actively nulling out RFI from unshielded arcing, co-channel and adjacent signals; the use of beamforming techniques and noise suppression.

- Wide deployable antennas on satellites for broadband and for use with hand-held mobile emergency telephones [13].

- Use of mini-HAPs (high-altitude platforms with buoyancy control) for high-data-rate communications. While it is unlikely that high-altitude platforms will be available over disaster zones to provide high-bandwidth communications because of the costs of the supporting balloon and launch, and aerial position maintenance, nevertheless, the costs are less than those for a satellite [14].

- Use of hand-held satellite telephones and power systems for immediate hazard assistance [15]. Cognitive radio for intelligent dynamic use of the frequency spectrum [16].

- Creative use of mobile frequencies in time of disaster: For instance, in smart phones an automatic weather or flood-warning system can be set to identify phone numbers of threatened areas, which can then receive an alert message sent from a prediction model.

3.2 Radio Propagation

Radio propagation (both terrestrial and satellite) may be affected by both the propagation environment in a disaster region and by aspects of the disaster (e.g., by firestorms [17], dust, sandstorms, foliage [18], and severe weather).

Meteorological conditions can significantly affect radio communications. In many disasters, the immediate effect on the local environment is to stimulate weather that can further affect the difficulties of rendering relief and assistance to the disaster zone. For example, dust and aerosol particles following earthquakes and volcanic activity may lead to rain and flooding. Consequently, the propagation medium within and surrounding a disaster zone will be a factor in providing effective data and information

Scope:	Standard: Conventional, robust, well-known applications.	Non-conventional, less familiar, maybe more modern applications.	New technology and / or combinations of techniques	Research environment: Potentially new concepts.
Limitations:	Potentially limited by familiarity, age, and experience	Less familiar, but still well understood.	Unfamiliar, currently not robust,	Pure research
Example:	Analogue HF radio	mobile phones	Left as an exercise for the reader.	(Maybe something you are doing right now!)
Approach: to increase value during disasters:	Tutorials, revised manuals	Professional reviews of techniques	Applied research, conference sessions	Research grant applications

Table 1. Radio-science contributions across a spectrum of inputs.

communications. Likely disaster scenarios can be defined, and their potential effects and practical improvements delineated, in preparation for emergency situations. Post-disaster case studies, describing the propagation conditions, will be valuable for future disaster scenarios.

Space-weather and ionospheric effects may further affect communications. This may seem an unlikely contributor, but in a disaster situation where HF communications are the best available medium, and this is already stressed, the expected space-weather conditions and their effects must still be accounted for in providing optimum operations. Examples include general HF frequency scheduling, impacts of the noise environment, the impact of ionospheric storms, and possible ionospheric degradation of VHF/UHF signals [19] and of SAR images, the potential for dilution of GPS accuracy [20], etc.

Communication propagation is dealt with mainly in URSI Commissions F and G.

3.3 Data Mapping and Analysis

Satellite imagery is now vital for warning of potential natural threats. By monitoring changes in land and water levels, mud and landslide potential may be assessed. Post-disaster, imagery is essential. For instance, it enables the assessment of damage in a disaster zone, the distribution of people and equipment on the ground, and can provide a temporal view of how these are changing. However, faster image processing is still needed to advance satellite topography and imagery [21].

Data (past and present) relevant to a disaster zone needs to be available and quickly accessible (especially terrain, underground services, weather conditions, water/sea currents). Mapping of data in real time, accessible storage, data analysis, and interoperability remain areas of research that are paramount for many uses, not least for disaster management. For disaster relief, authorities need to be able to access data remotely, and to map data that is brought out of the disaster zone in near real time.

While URSI does not provide data advice, it is clearly linked to the provision of the data, both through communications, and also through understanding the processes leading to its generation.

3.4 Remote Sensors

Multi-spectral processing offers composite analysis of a specific situation (e.g., heat, viral, radiation). THz bio-imaging and screening [22] should be added to this item of study. Lying between electronics and optics on the spectrum, these techniques offer many applications for future management of disaster situations. THz applications include Earth remote sensing, chemical analysis, and explosive detection, among others.

Ultra-wideband (UWB) pulsed radar devices have the potential to locate people and assets under rubble. Ground-penetrating radar [23], UWB radar, and THz imaging have the potential to reveal information beyond some building structures and under clothing materials.

SAR (synthetic-aperture radar) can detect topographical changes relating to potential natural hazards, although this is usually in non-real time. HF skywave and surface-wave radar can have applications in sea-surface monitoring. Remote sensing of the Earth and the atmosphere, with an ever-expanding collection of advanced instruments, has great potential for monitoring our planet, and for providing both warnings and relief in disasters: both natural and manmade. The development of these products depends on scientific analysis and modeling of the radiowave interactions with the environment. For emergency situations, it is essential to have standard connectivity, interoperability, and access for sensors (e.g., video, radio, radiation, heat, temperature) into portable communication systems. Although this is not exactly radio science – and, hence, URSI – there is an opportunity to work towards new systems of data transmission (e.g. radio/optical) that can operate with all sensor types and communication units.

Robotic transports, and their sensors, used in dangerous situations need secure and reliable communications. URSI might support aspects of these engineering problems.

There is an increasing use of UAVs (unmanned aerial vehicles) over a disaster zone because of their obvious advantages in high-resolution monitoring of the situation [24]. UAV control over a reasonable distance, possibly under difficult environmental conditions, is an area for increased research. Can they be operated in such a way that they will not be a danger to other aircraft in the area? Is hijacking a potential problem during a disaster?

URSI either spans these areas of interest, or has peripheral connections to them. There are many radio-science opportunities in the area of remote sensing.

3.5 Signal Processing

Electromagnetic compatibility (EMC) and RF interference reduction through signal processing remain critical areas for study, especially where an emergency creates a focus for hugely increased radio usage and, hence, incompatibility between systems.

Discrimination of weak wanted radio signals from radio-frequency interference (RFI) has been improved through development of efficient algorithms and techniques, but can be further developed, and the use of these techniques increased with increasing familiarity. Naturally, case studies during different disaster environments form an important test bed for further research.

Under extreme conditions, during a disaster, it is likely that more information will need to be sought than during normal times. Consequently, increased bandwidth will be needed to provide high-resolution video images, for instance. Research leading to greater data communication flows will be especially useful.

3.6 Static Information Packages

Pre- and post-disaster, information packages (possibly Web-based) can explain potential disaster scenarios, their background and contexts (e.g., volcanic dust [25], tsunami generation, landslides, sandstorms, fire storms, solar storms, tornadoes, sea-wave generation, earthquakes, oil explosions, and radiation exposure).

URSI has the competence and skills to implement a number of static information sites relating to a standard set of natural and manmade disasters. At times of specific incidents, URSI could provide updated specialist information available online so that the public and/or emergency managers could better understand the context of the situation.

3.7 Further Radio-Science Examples

The previous sections provided several radio-science examples that could assist in managing a disaster. Below, further highlights are identified, some or all of which, if deployed, could save lives. The list is not definitive, but provides further examples of the many ways radio science might contribute to lessening the impact of a disaster.

- UWB ground-penetrating radar and THz imaging to assist in probing structures
- Doppler radar for movement detection (structures, mining, earthquake, landslides)
- RF and wireless ID tagging of equipment for swift collation in disaster situation
- Emergency communications via mobile phones; preferably, satellite phones
- Emergency beacons included in phones
- Monitoring, mapping, and predicting sea-wave paths across oceans
- Weather and Earth-observation satellites
- Communication satellites (Earth-space radio paths)
- Satellite-based positioning
- Earth movement and ground deformation detection (Doppler radar, visual, SAR)
- Earth pressure and stress changes prior to slips/earthquakes
- Satellite detection of water levels, volcanic activity, wind direction/intensity, sea temperatures
- Database of radio-propagation characteristics expected in disaster zones (e.g., dust, meteorological, sand, firestorm, foliage, building structures)
- High-resolution sensors to pinpoint radio, audio, video signals from individuals
- Multi-spectral images for gauging water depth, vegetation coverage, soil moisture content, and presence of fires
- Noise reduction in disaster scenarios (reducing radio noise, interference, and spurious data)
- Integrated network planning (sensor grid to collect information, command and control grid for assessment and decision making, engagement grid that directs action, and information grid providing the network infrastructure)
- Telemedicine: remote participation in and/or control of medical procedures in the immediate disaster zone
- Use of radio interference and SAR to identify Earth movements (earthquakes, landslides), and for accurate topography

The plan is for this document to form the basis for the WG1 Web site, which over time will grow in content and value. Not all the subjects noted in these sections have suitable references. Hopefully, readers recognizing the work will supply references (both from the literature and on the Web), so this summary can be a more-complete reference.

4. Summary

An example often quoted is that despite the continuing increase in aviation, fewer fatal accidents occur because of constantly improving technology, particularly radio communications, satellite positioning, and instrument monitoring. Increased global coordination of observation systems and data sharing is undoubtedly helping to reduce losses in disasters, despite the increasing population.

Radio-science innovations have given rise to many remarkable advances in radio communications, radar, and associated technologies that have reduced the risks of natural and human-induced disasters. This paper proposes that in addition to having made significant contributions to disaster mitigation in the past, there are radio-science studies that can improve the management of potential disasters even further in the future.

It is clear URSI can play an important role in disaster relief. URSI Inter-Commission Working Group 1 on Natural and Human-induced Hazards and Disasters is scoping the work being carried out within the URSI community, with the intention of bringing it to the attention of emergency managers through URSI channels. All Commissions are capable of improving this list of studies, and all URSI members are encouraged to communicate their ideas to both their Commission Chairs and to the WG 1 Chair, Phil Wilkinson (e-mail: phil@ips.gov.au).

There are no doubt many untapped opportunities where radio science can further contribute. Don't underestimate the knowledge you have: sharing it could make all the difference in a disaster.

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CONFERENCE REPORTS

COSPAR 2010 – C4.1: Representation of the Auroral and Polar Ionosphere in IRI

Bremen, Germany, July 18 -20, 2010

A two-day session on the 'Representation of the Auroral and Polar Ionosphere in the International Reference Ionosphere (IRI)' was held during the 38th Scientific Assembly of the Committee on Space Research (COSPAR) in Bremen, Germany. IRI is the internationally recommended empirical model for the ionosphere. It is the result of a joint project of COSPAR and the International Union of Radio Science (URSI), and the general assemblies of these two organizations are the main venue for discussions of model performance, shortcomings, improvements, additions, applications and most importantly decisions regarding the next version of the model. These meetings are also the platform for initiating collaborative projects with the goal of future improvements of the model.

A total of 42 presentations were given which were grouped into 5 topical areas: IRI at High Latitudes, GNSS Observations and IRI, Representation of the Topside Ionosphere in IRI, Improving the Description of Solar Forcing in IRI, and New Inputs to IRI. A hallmark of IRI sessions is the wide variety of data sources used to check and improve the model. The Bremen meeting was no exception and included presentation that were based on satellite measurements from TIMED, TOPEX, Jason, GPS, COSMIC, CHAMP, Alouette, ISIS, ACTIVE, APEX, CORONOS-I, AE-E, AE-E, and OGO-5, and on ground based measurements from the global network of ionosondes, and on incoherent scatter radar observations from EISCAT, Kharkov, and Arecibo.

The majority of presentations were focused on the performance of the IRI model at high latitudes and possible improvements. A first step towards this goal is the introduction of auroral boundaries in the next version of the model, IRI-2010, based on the model developed by Yongliang Zhang and Larry Paxton (JHU/APL, USA) using TIMED-GUVI data. The next step now will be the representation of typical auroral characteristics like the density trough and temperature cusp. Another TIMED instrument, SABER, has provided Chris Mertens (NASA Langley, USA) with the database to develop an auroral E-region storm model which is also scheduled for inclusion in IRI-2010. Several years of continuous EISCAT measurements during the recent deep and extent solar minimum are a unique data source that has not yet been exploited for IRI modeling at high latitudes.

One of the most anticipated IRI improvements is the new Neural Network model for the F-peak critical frequency (foF2) and for the propagation factor M(3000)F2 that is related to the F-peak height (hmF2). These models were developed by Lee-Anne McKinnell (HMO, South Africa) and Elija Oyeyemi (University Lagos, Nigeria) based on a large volume of global ionosonde measurements. A presentation during this meeting showed results from a high latitude validation of the new foF2 model and revealed improvements of between 20% and 60% versus the currently used CCIR and URSI foF2 models. This is as a result of new high latitude ionosonde data being made available for inclusion in this model.

GPS and COSMIC data are a rich data source for improving the global reliability of the IRI model and several presentations during this meeting reported on data-model comparisons and on new models developed based on these data sources. Bias factors and inherent limitations of tomographic and occultation methods during times of rapid change (sunrise, sunset, storms) and in regions of steep gradients (Equatorial Anomaly region) have to be kept in mind when using these data for global modeling. But GNSS measurements are an excellent data source for the Total Electron Content (TEC) and a number of methods now exists to turn the combined global GNSS measurements (high and low GPS, TOPEX, Jason) into a global mapping of TEC; several methods were discussed during this meeting (Andrzej Krankowski, U Warmia and Marzury, Olsztyn, Poland; Denise Dettmering, German Geodetic Institute, Munich, Germany; Mahdi Alizadeh Elizei, TU Vienna, Austria).

A new model for the electron temperature was presented by Vladimir Truhlik (IAP, Prague, Czech Republic). This model was developed with the help of a large database of in situ measurements from many satellites covering more than 2 solar cycles. The significant improvement over the current IRI electron temperature model is the explicit inclusion of variations with solar activity in this new model. It will be included in IRI-2010.

Additionally, in IRI-2010 the global representation of the electron and ion density in the region below the F-peak is significantly improved based on the modeling work of David Altadill (Ebro Observatory, Spain) and Phil Richards (GMU, USA) that had been reported at earlier IRI workshops.

Presentations from this session will be considered for a special issue of Advances in Space Research. Papers from the previous COSPAR IRI session in Montreal, Canada in 2008 have just been published as Issue 8 in Volume 46 of Advances in Space Research. The term of office for COSPAR officials is limited to 8 years and the current team chairing the IRI Working Group consisting of Chair, Bodo Reinisch (UML, USA), and his two deputies, Lida Triskova (IAP, Czech Republic; Vice-chair for URSI) and Martin Friedrich (TU Graz, Austria; Vice-chair for COSPAR) had now reached this limit. Therefore election of a new leadership team was an important point on the agenda for the IRI business meeting. A new team was proposed and elected unanimously: Lee-Anne McKinnell (HMO, South Africa; Chair), Shigeto Watanabe (Hokkaido U, Japan; Vice-Chair for COSPAR), Vladmir Truhlik (IAP, Czech Republic; Vice-Chair for URSI). Three new members were elected into the IRI Working Group: Claudia Stolle (National Space Institute, Denmark), Ivan Galkin (UML, USA), and Hanna Rothkaehl (Space Research Center, Poland). The next IRI Workshop is scheduled for 10-14 October 2011 at the Hermanus Magnetic Observatory in Hermanus, South Africa. For the 2013 Workshop the IRI community has been invited to hold its meeting at the University of Warmia and Mazury in Olsztyn, Poland.

EMTS 2010 - Commission B International Symposium on Electromagnetic Theory

Berlin, Germany, August 16 -19, 2010

The International U.R.S.I. Commission B „Fields and Waves“ holds an international triennial symposium on Electromagnetic Theory (EMTS) starting in 1953 in Montreal via the events 2004 in Pisa/Italy and 2007 in Ottawa/Canada to Berlin/Germany in 2010, and further to Hiroshima/Japan in 2013.

The Berlin symposium can be considered as particularly successful; it was held in the Steigenberger Hotel in the former western part of Berlin that found appreciation by practically all participants not only for technical organization but also for catering during breaks and lunch. The symposium attracted about 260 participants during the week August 16 to 19. The scientific program was set up by a Technical Program Committee consisting of the national Commission B Chairs supported by a Technical Advisory board formed by well-known radio scientists.

Topics included the conventional topis of the Commission B terms of reference as well as special sessions proposed by respective conveners. A memorial session for the late Femke Olyslager was organized by Daniel De Zutter and Ludger Klinkenbusch. All accepted and presented papers, reviewed by at least two referees, will be published by IEEEXplore. In addition, a Special Section of the Journal „Radio Science“ consisting of selected papers recommended by conveners and/or session chairs is in preparation. The local arrangements for the symposium including preparation of the abstract booklet and the proceedings, and maintenance of the website was

Figure 1: 21 Young Scientist Awardees received a certificate during the Banquet.

in the hands of the Local Organizing Committee headed by Ludger Klinkenbusch. A highlight of the symposium was by no doubts the Banquet that was held in the glass courtyard (Glashof) of the Jewish Museum.

U.R.S.I. symposia carefully maintain the tradition that is also true for the General Assemblies to support Young Scientists. The Berlin symposium received 49 applications satisfying the criteria that the applicant must be the first author of an accepted paper and that he must be younger than 35. Consuetudinarily, for the Commission B symposia the Vice-Chair of the Commission, presently Giuliano Manara, is responsible to handle the program. With the help of reviewers he proposed 21 nominees, and, as a matter of fact, due to the financial support of U.R.S.I. and several sponsors, all 21 could be supported meaning that their registration fee was waived, and that they received a perdiem and an invitation to the Banquet; those from developing countries also received a financial travel support. Certificates honoring the Young Scientist Awardees were issued by Giuliano Manara during the Banquet. (see Figure 1)

Two more sponsors – the software company CST and the consulting company NavCom – arranged for a Young Scientist Best Paper Award: first prize Euro 1000, second prize Euro 750, third prize Euro 500. To select potential awardees from the Young Scientists reviewers proposed 6 nominees who, in addition to their oral paper, had to prepare a poster that was evaluated by a selection committee.

Finally, the awardees were announced during the Banquet: 1. Yan Kaganovsky from Israel, 2. Alireza Kazemzadeh from Sweden, 3. Giorgio Carluccio from Italy. All three awardees received a respective certificate and, of course, the prize money.

Karl J. Langenberg
Chair of International U.R.S.I. Commission B
Chair of the Technical Program Committee

ISRSSP 2010 - SEcond International Symposium on Radio Systems and Space Plasma

Sofia, Bulgaria, August 25 -27, 2010

The Second International Symposium on Radio Systems and Space Plasma (ISRSSP 2010) was held in Hilton Sofia hotel (Bulgaria) on August 25-27. The symposium was organized by the Interdisciplinary Institute for Collaboration and Research on Enterprise Systems and Technology - IICREST, under the auspices of the Union of Radio Science - URSI, and in collaboration with the Institute for Systems and Technologies of Information, Control and Communication – INSTICC, the Bulgarian Academy of Sciences, and Technical University – Sofia. Among the technical co-sponsors were the Municipality of Sofia and Ruse University. There were 50 participants from 11 different countries. The ISRSSP'10 proceedings can be downloaded at: http://www.isrssp.org/IS-RSSP/Documents/ISRSSP_2010_Proceedings.pdf

During the opening session of the symposium, the ISRSSP'10 Chair, Prof. Blagovest Shishkov and the President of URSI, Prof. François Lefeuvre addressed the audience with speeches. In addition, the ISRSSP'10 participants were addressed by the Mayor of Sofia Municipality, Mrs. Yordanka Fandakova, the President of the Bulgarian Academy of Sciences, Prof. Nicola Sabotinov, and the Rector of Technical University – Sofia, Prof. Kamen Veselinov.

The program of ISRSSP'10 included 43 presentations, among which 39 presentations reflecting research findings reported in scientific papers. All scientific papers were published in the ISRSSP'10 proceedings, including 30 invited papers and 9 regular papers.

With regard to the mentioned 43 presentations, 2 were from Austria, 9 from Bulgaria, 1 from Brazil, 5 from France, 1 from Italy, 15 from Japan, 1 from Poland, 1 from Romania, 2 from Russia, 1 from Turkey, and 2 from USA. All presenters are distinguished scientists from the mentioned countries.

The selected papers are a good illustration of different relevant topics that are currently under research, presenting original contributions. Among the topics considered in the papers are: Intelligent Methods of Radio-Communication Systems and Signal Processing, Transionospheric Propagation and Satellite Observations, Methods for Analyzing Non-linear Interactions of Space Plasma, Radio Science Aspects of SPS Systems. These topics are in line with the ISRSSP'10 agenda (discussed below).

Covering the topics of URSI commissions C, G, H (toward space plasma) and Space Solar Power Systems (SSPS) and being influenced by all these commissions, the ISRSSP'10 scientific program was concerned with the international development and application of the following

range of aspects:

(i) Radio-Communication (RC) Systems and Signal Processing
(ii) Transionospheric propagation; Investigations of Space Environments via Satellite Observations
(iii) Generation and Propagation of Waves in Plasma. Interaction between Waves and Wave Particles
(iv) Solar Power Satellite (SPS) Systems and related Radio Technologies; Further Directions in SPS Systems

It is not only that ISRSSP'10 adopted the URSI research agenda but it has as well spread its influence to a broader audience including all those researchers and practitioners interested in the topics of the symposium. Their participation has additionally strengthened the URSI-related research and also the RC research developments. All this inspired the organizers to go forward with the organization of ISRSSP'13, contributing to the dissemination of radio science -related knowledge.

The ISRSSP'10 program included not only the presentation of the papers that represent the main content of the ISRSSP'10 proceedings but also discussions that concern the further ISRSSP-URSI agenda. These high points in the symposium program have definitely contributed to the reinforcement of the URSI-related research and knowledge dissemination.

The preparation of this symposium required the dedicated effort of many people. We must thank the authors, whose research and development efforts are recorded in the ISRSSP'10 proceedings. We thank also the members of the program committee for their support and expertise. A special thanks to the President of URSI – Prof. Francois Lefeuvre whose collaboration was fundamental for the success of this symposium. Last but not least, we would like with great pleasure to thank IICREST for the brilliant organization.

And in the end, considering the further ISRSSP-USRI developments, an URSI-IICREST Executive Meeting was held in Sofia on the 25th of August, attended by the President of URSI, the President of URSI – Bulgaria, the Chairman of IICREST, and key representatives of URSI Bulgaria and URSI Japan. Following a discussion, the present URSI-IICREST members unanimously decided to invite IICREST to organize in close collaboration with URSI the 3rd International Symposium on Radio Systems and Space Plasma (ISRSSP 2013) to take place in Sofia (Bulgaria) on 28-30 August, 2013.

Blagovest Shishkov
Chair of ISRSSP'10
President of the Bulgarian URSI Committee

4th VERSIM Workshop

Prague, September 13 -27, 2010

General information

The 4th workshop of the URSI/IAGA Joint Working Group on VLF/ELF Remote Sensing of the Ionosphere and Magnetosphere (VERSIM) took place in Prague, Czech Republic, on 13-17 September 2010. The workshop was organized by the Institute of Atmospheric Physics, Prague, Czech Republic and by Charles University, Faculty of Mathematics and Physics, Prague, Czech Republic. The scientific sponsorship and financial support for this workshop has been provided by the International Association of Geomagnetism and Aeronomy (IAGA) and by the Union Radio-Scientifique Internationale (URSI). More details can be found on the workshop website (http://www.ufa.cas.cz/versim10/).

The Scientific committee consisted of János Lichtenberger (Eötvös University, Budapest, Hungary), Jacob Bortnik (University of California, Los Angeles, USA), Jyrki Manninen (Sodankyla Geophysical Observatory, Sodankyla, Finland), Craig J. Rodger (University of Otago, Dunedin, New Zealand), Yoshiharu Omura (Kyoto University, Kyoto, Japan), David Shklyar (Institute of Space Research, Moscow, Russia), Michel Parrot (Laboratoire de Physique et Chimie de l'Environnement et de l'Espace, Orleans, France), Ondřej Santolík (Institute of Atmospheric Physics and Charles University, Prague, Czech Republic)

The Local organizing committee was composed of the (Institute of Atmospheric Physics and/or Charles University, Prague, Czech Republic) Jaroslav Chum, Andrea

Fig 1: Participants of the 4th VERSIM workshop. Photo taken on 16 September 2010 by A. Collier

Saglová, Jan Souček, František Hruška, František Jiříček, Pavel Tříska, Benjamin Grison, Mykhaylo Hayosh, Ondřej Santolík, Jiří Pavlů andJana Šafránková

Topics

. ELF/VLF waves as diagnostic tools for key ionospheric and magnetospheric physical parameters. Wave propagation in magnetosphere and ionosphere and in Earth-ionosphere waveguide. Plasma structures and boundaries - morphology and dynamics.
. Triggered emissions and nonlinear wave phenomena. Wave-particle and nonlinear wave-wave interactions.
. ELF/VLF waves and the radiation belts - particle acceleration and loss. Wave-induced particle precipitation.
. Effects of ELF/VLF radiation on the atmosphere. Sprites and the effects of lightning on the ionosphere. Schumann resonances, world wide lightning activity and climate changes
. Instrumentation

Abstracts and sessions

47 abstracts were received. They are listed online on the workshop website (http://www.ufa.cas.cz/versim10/list-prg.php), together with an alphabetical list of authors (http://www.ufa.cas.cz/versim10/list-author-index.php). The scientific committee organized these abstracts in 7 oral sessions and one poster session.

. Session I – Triggered emissions and wave-particle interactions (1 invited + 4 contributed talks)
. Session II – Chorus and quasi-periodic emissions (5 contributed talks)
. Session III – Whistlers (7 contributed talks)
. Session IV – VLF data sets and campaigns (1 invited + 5 contributed talks)
. Session V – DEMETER (6 contributed talks)
. Session VI – Subionospheric VLF propagation (4 contributed talks)
. Session VII – Ionoshere, Martian ionosphere, meteors (4 contributed talks)
. Poster session (8 posters)

VERSIM business meeting has been also held as in the frame of the workshop. The workshop program and abstract book can be downloaded from http://www.ufa.cas.cz/versim10/program.pdf

Participants

The workshop attracted 40 participants from 15 countries, ranging from New Zealand to Russia (ordered by latitude) including 9 students and young scientists under 35 from 5 countries. The final list of participants from each country can be downloaded from http://www.ufa.cas.cz/versim10/participants_final.pdf. The updated workshop program including all last-minute changes can be downloaded from http://www.ufa.cas.cz/versim10/program_final.pdf

Scientific Highlights

VERSIM workshops are now a strong feature of the VERSIM community, which were well supported by the membership and included many very strong presentations. In his invited talk, Yoshiharu Omura described an exciting theory and simulations of VLF triggered emissions. His talk was followed by a session containing theoretical and experimental papers on triggered emissions and chorus with clear implications for the currently high-profile subject of acceleration and losses of energetic particles in the Earth's radiation belts. Another exciting invited talk was given by Andy Smith on extending continuous quantitative measurements of ELF/VLF noise at the Halley station in Antarctica.

Generally, much progress there has been since the first workshop in 2004. For example, data from the DEMETER satellite, which was newly launched in 2004, is now being used in a very large percentage of the talks at the 4th workshop (including a dedicated session). Janos Lichtenberger reported on the auto-detection of tens of millions of whistlers, and his successes in developing an automatic whistler scaling technique. SAVNET (South American VLF Network) was being used to study solar flares and gamma ray bursts occurring thousands of light years distant. Another example is the WWLLN lightning location network. Craig Rodger reported that this network has over 50 operating stations, reporting 115.4 million lightning strokes in 2009. Many of the talks presented in this workshop used WWLLN data. Through the VERSIM workshops there is increasing collaboration in our community, strengthening the scientific quality and overall "force" of the community.

URSI Financial Support

Commission H of URSI decided to sponsor the 4th VERSIM workshop with an amount of EUR 1300. The money was used by the local organizing committee of the 4th VERSIM workshop to partially support young scientist and students under 35 years of age at the time of the meeting and to partially support other deserving participants judged by the local organizing committee of the 4th VERSIM workshop to be in need of financial support.

Next workshop

It has been widely accepted that following the strengths of our previous VERSIM workshops, there would be real value in a 5th VERSIM workshop. At this point it seems sensible to have this in 2012, as neither IAGA or URSI will meet that year. Jean-Pierre Raulin (Brazil) has invited the VERSIM community to meet in the state of Sao Paulo, Brazil in September 2012. He gave a presentation on this concept – at this point the suggestion is to hold the meeting in a hotel ~60 km away from the city of Sao Paulo, in a comparatively quiet region near to their main observatory site. The observatory site would provide a potential excursion during the meeting, and also has student-level accommodation for approximately 12 people. Another excursion would be to an INPE space research centre some 1 hours drive away. Jean-Pierre outlined the various Brazilian organisations who would be expected to support the meeting (logistics, administration and financial support), and indicated that he hoped quite a few participants could be assisted to attend – considerably more than in previous VERSIM workshops.

ISSSE 2010

Nanjing, China, September 12 -20, 2010

2010 International Symposium on Signals, Systems and Electronics (ISSSE2010) had been held in Mandarin Garden Hotel, Nanjing, China on September 17-20, 2010. This symposium is held every three years, and is organized under the guidance and with sponsorship of the international steering committee of the URSI Commission C (Radiocommunication Systems and Signal Processing) and D (Electronics and Photonics).

The ISSSE 2010 was co-sponsored by the Southeast University (SEU), URSI Commission C and URSI Commission D and was Technically co-sponsored by the Nanjing University of Science and Technology (NUST), the Nanjing University, the PLA University of Science and Technology (PLAUST), the Nanjing University of Aeronautics and Astronautics (NUAA), the Nanjing University of Posts & Telecommunications (NUPT), Nanjing Purple Mountain Astronomical Observatory, IEEE MTT-S, the Antenna Society of Chinese Institute of Electronics (CIE), the Microwave Society of CIE, IEEE Nanjing Section, IEEE MTT-S/AP-S/EMC-S Joint Nanjing Chapter, IEEE Beijing Section, IEEE Shanghai Section, IEEE Harbin Section and IEEE Chengdu Section

There were 310 papers submitted to ISSSE 2010 from 33 countries and regions. 253 papers were accepted after peer reviewed by TPC Members. The acceptance rate of the conference is 82%. The registered attendee was 255 including 245 pre-registration and 10 on-site registration. The conference events consist of 5 Keynote Presentations, 18 oral sessions, 7 poster sessions and 3 shared tutorials with ICUWB2010.

Figure 1. The Conference was held in the Mandarin Garden Hotel

Figure 2 : the Key Committee Members

Figure 1: Opening Ceremony

Figure 2: One of the Keynote Speeches

Prof. Takashi Ohira, Chairman of URSI Commission C, Toyohashi University of Technology, Japan attended and registered ISSSE2010 as URSI Representative. He also attended the ISC meeting held on Sept.19, 2010.

Best Student Paper Awards

- **A High Linearity Reconfigurable Down-Conversion Mixer for Dual-Band Applications** Tao Li, Fengyi Huang, Yan Wang and Xinrong Hu, Southeast University

- **Improvement the Q-factor of Multi-Band Inductor with 90 μm Silicon Substrate on Plastic** H. M. Chang, H. L. Kao, S. P. Shih, Y. C. Lee, C. Y. Ke, L. C. Chang, C. H. Wu, Jeffrey S. Fu and Nemai C. Karmakar, Chung Hua University

- **Novel Folded Single Split Ring Resonator and Its application to Eliminate Scan Blindness in Infinite Phased Array** Mingchun Tang, Shaoqiu Xiao, Tianwei Deng and Bingzhong Wang University of Electronic Science and Technology of China

Figure 3: Best Student Paper Competition

There were one platinum sponsor and four exhibitors attending ISSSE2010, which are as followed:

- Agilent Technologies (Platinum Sponsor), Email: www.agilent.com.cn
- Farran Technology (Exhibitor), Email: www.farran.com
- Rohde & Schwarz China Ltd. (Exhibitor), Email: www.rohde-schwarz.com.cn
- Shanghai Huaxiang Computer Communication Engineering Co., Ltd (Exhibitor), Email: www.shx-sh.com
- Isola Group (Exhibitor), Email: www.isola-group.com

Figure 4: Industry Exhibition

4TH IAGA/ICMA/CAWSES II TG4 Workshop on Vertical Coupling in the Atmosphere-Ionosphere System

Prague, Czech Republic, 14 - 18 February 2011

Conference description

The Earth's atmospheric regions are intricately coupled to one another via various dynamical, chemical, and electrodynamic processes. However, the manner in which the couplings take place due to varying energy inputs from the Sun and from the lower atmosphere is a question that is yet to be understood. The coupled effects can be in terms of the modulation of the waves from lower to upper atmosphere as well as from low-to high-altitudes, electrodynamic and compositional changes, and plasma irregularities at different latitudinal regions of the globe due to the varying energy inputs. The MLT region is a critical region in the coupling between the lower/middle atmosphere and the upper atmosphere/ionosphere since it is here that physical processes filter and shape the flux of waves ascending through the mesosphere into the overlying thermosphere. On the other hand it is reasonable to presume that there might be a link between solar variability and the changes in the middle atmosphere and climate variables. This requires much improved knowledge and understanding of the solar effects on the coupling processes. This workshop solicits papers dealing with experiments, observations, modelling, and data analysis that describe the effects of atmospheric coupling processes within the atmosphere-ionosphere system. Recent results from the campaigns of CAWSES-II program are particularly welcome. It will address both theoretical and empirical recent results concerning the coupling mechanisms through dynamics, composition and electrodynamics. The workshop will be particularly focused on the dependence of coupling processes on the solar and geomagnetic activity, the downward control effects transferring from the strongly solar dependent structure to the lower atmospheric levels.

Contact

Dr. Petra Koucka knizova
Institute of Atmospheric Physics
Bocni II/1401, 14131
Praha 4, Czech republic
Fax: +420 272 763 745
E-mail: pkn@ufa.cas.cz
http://www.ufa.cas.cz/html/conferences/workshop_2011/

Abstract submission deadline: 5 December

13TH INTERNATIONAL SYMPOSIUM ON MICROWAVE AND OPTICAL TECHNOLOGY - ISMOT 2011

Prague, Czech Republic, 20 - 23 June 2011

ISMOT-2011 will be held in Prague on June 20 - 23, 2011. The venue of the meeting will be hotel Olympic. The research fields covered by ISMOT are in the area of microwave components and circuits, optical components, microwave and optical communication systems, electromagnetic theory, antennas, microwave photonics, and all other topics of interest to microwave and optical communities.

The main objective of ISMOT is to offer an international forum for the exchange of new ideas, thoughts, and realizations on physics, technologies, and applications of microwave, optoelectronics, and related fields in industry, communications, science, research, medicine etc.

Selected papers of ISMOT-2011 will be published in the International Journal ofMicrowave and Optical Technology (IJMOT) as full-length papers.

General Chair

Prof. B.S. Rawat
University of Nevada, Reno, NV, USA

Symposium Chair

Prof. Jan Vrba
Czech Technical University, Czech Republic

International Advisory Committee

Prof. B.S. Rawat, USA, Chair
Prof. Jan Vrba, Czech Republic, Vice-Chair
Prof. R.S. Gupta, INDIA, Vice-Chair

Important dates

Deadline for abstract submission: 15 February, 2011
Notification of accepted abstracts: 10 March, 2011
Early Bird Registration deadline: 31 March, 2011

The up-dated information about paper format, submission method etc. is available at: http://www.ismot2011.com

Conference Topics

1. Microwave components and systems
2. Millimeter-wave components and systems
3. Solid-state devices & Circuits
4. MICs/MMICs
5. Microwave materials
6. Microwave superconductivity applications
7. Communication systems
8. Antennas and radar technologies
9. Numerical methods and CAD techniques
10. Propagation/scattering and measurements
11. Electromagnetic theory
12. Optoelectronics
13. Optical fibers and waveguides
14. Optical solutions
15. Optical communications/network and sensors
16. Optical multiplexing/demultiplexing
17. Laser technology
18. Metamaterials in microwave and photonics
19. Microwave photonics
20. Remote sensing
21. Signal processing
22. Biological effects and medical applications
23. Industry and environmental effects
24. Microwave/optical education
25. EMI/EMC
26. Embedded Systems
27. Satellite & Wireless Communication Systems
28. MEMS
29. Packing, Interconnects & MCMs
30. Any other relevant topics

Conference Secretariat

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URSI CONFERENCE CALENDAR

An up-to-date version of this Conference Calendar, with links to various conference web sites can be found at http://www.ursi.org/en/events.asp

February 2011

10th International Congress of the European Bioelectromagnetic Association
Rome, Italy, 21-24 February 2011
Contact : Dr. Micaela Liberti, Department of Information Engineering, Electronics and Telecommunications, “La Sapienza” University of Rome, Via Eudossiana 18, I-00184 Roma, Italy, Fax: +39 06 44585918, E-mail: Chair@ebea2011.org

4th IAGA/ICMA/CAWSES-II Workshop: Vertical Coupling in the Atmosphere-Ionosphere System
Prague, Czech Republic, 14-18 February 2011
cf. Announcement in the Radio Science Bulletin of December 2010, p. 59.
Contact : E-mail: workshop_2011@ufa.cas.cz, Website: http://www.ufa.cas.cz/html/conferences/workshop_2011/

March 2011

12th International Com F Triennial Open Symposium on Radio Wave Propagation and Remote Sensing
Garmisch-Partenkirchen, Germany, 8-11 March 2011
cf. Announcement in the Radio Science Bulletin of September 2010, p. 76-77.
Contact: Prof. M. Chandra, Microwave Engineering and Photonics, Electrical Eng & Information, T.U. Chemnitz, Reichenhainer Strasse 70, D-09126 CHEMNITZ, GERMANY, Fax : +49 371 531 24349, E-mail : madhu.chandra@etit.tu-chemnitz.de

April 2011

JURSE 2011 - Joint Urban Remote Sensing Event 2011 (formerly URBAN)
Munich, Germany, 11-13 April 2011
Contact: Photogrammetry & Remote Sensing, Technische Universitaet Muenchen, Arcisstr. 21, D-80333 München, Germany, Fax: +49 89 2809573, E-mail: pf at bv.tum.de, Web: http://www.pf.bv.tum.de/jurse2011/

CEM 2011 - Eight International Conference on Computation in Electromagnetics
Wroclaw, Poland, 11-14 April 2011
Contact: Prof. Jan K. Sykulski, Chairman of CEM 2011, Electrical Power Engineering Research Group, School of Electronics and Computer Science, University of Southampton, Southampton, SO17 1BJ United Kingdom, Fax +44 23-8059 3709, E-mail: jks at soton.ac.uk, Web: http:// www.cem2011.com

June 2011

ISMOT 2011 - 13th International Symposium on Microwave and Optical Technology
Prague, Czech Republic, 20 - 23 June 2011
cf. Announcement in the Radio Science Bulletin of December 2010, p. 59-60.
Contact: Prof. Jan Vrba, Czech Technical University in Prague, Dept. of Electromagnetic Field, CVUT - FEL, K 13117, Technicka 2, 166 27 Prague 6, Czech Republic, Email: vrba@fel.cvut.cz, Website: http://www.ismot2011.com

July 2011

IconSpace 2011: International Conference on Space Science & Communication
Genting Highlands, Pahang, Malaysia, 12-13 July 2011
Contact: Dr. Wayan Suparta, Institute of Space Science (ANGKASA), Level 2, Faculty of Engineering and Built Environment, Universiti Kebangsaan Malaysia, 43600 UKM Bangi, Selangor, Malaysia, E-mail: wayan@ukm.my, w_suparta@yahoo.com, Website: http://www.ukm.my/ispace

August 2011

XXXth URSI General Assembly and Scientific Symposium
Istanbul, Turkey, 13-20 August 2011
Contact: URSI Secretariat, c/o INTEC, Sint-Pietersnieuwstraat 41, B-9000 Gent, Belgium, Fax +32 9 264 4288, E-mail: info@ursi.org and ursigass2011@ursigass2011.org , http://www.ursigass2011.org/

September 2011

ICEAA-APWC 2011 - International Conference on Electromagnetics in Advanced Applications
Torino, Italy, 12-17 September 2011
cf. Announcement in the Radio Science Bulletin of September 2010, p. 77-78.
Contact: Prof. P.L.E. Uslenghi, Dept. of ECE (MC 154), University of Illinois at Chicago, 851 S. Morgan Street, CHICAGO, IL 60607-7053, USA, Tel : +1 312 996-6059, Fax : +1 312 996 8664, E-mail : uslenghi@uic.edu , http://www.iceaa.net

EMC Europe 2011
York, United Kingdom, 26-30 September 2011
Contact: Prof. A. C. MARVIN, Department of Electronics, University of York, Heslington, YORK, YO10 5DD, UNITED KINGDOM, Phone : +44 (0)1904 432342, Fax : +44 (0)1904 433224, E-mail : conference@emceurope2011.york.ac.uk, website : http://www.emceurope2011.york.ac.uk/

October 2011

ISAP 2011 - 2011 International Symposium on Antennas and Propagation
Jeju, Japan, 25-28 October 2011
cf. Announcement in the Radio Science Bulletin of June 2010, p. 53-54.
Contact: 5F Daehan Bldg., #1018 Dunsan-Dong, Seo-Gu, Daejeon 302-120, Korea, Tel : +82-42-472-7463, Fax : +82-42-472-7459, isap@isap2011.org, *http://*www.isap2011.org

July 2012

COSPAR 2012 - 39th Scientific Assembly of the Committee on Space Research (COSPAR) and Associated Events
Mysore, India, 14 - 22 July 2012
cf. Announcement in the Radio Science Bulletin of September 2010, p. 78-79.
Contact: COSPAR Secretariat, c/o CNES, 2 place Maurice Quentin, 75039 Paris Cedex 01, France, Fax: +33 1 44 76 74 37, cospar@cosparhq.cnes.fr , http://www.cospar-assembly.org/

URSI cannot be held responsible for any errors contained in this list of meetings.

News from the URSI Community

NEWS FROM A MEMBER COMMITTEE

INDIA

National Workshop on Atmospheric and space sciences (NWASS 2010)

NWAS 2010 was held from 23-24 November, 2010, in the Meghnad Saha Auditorium, Rashbehari Siksha Prangan (Rajabazar Science College) in Kolkata.

It was organized by S. K. Mitra Centre for Research in Space Environment, the Institute of Radio Physics and Electronics, UGC Networking Resource Centre in Physical Sciences and the University of Calcutta.

This workshop received financial support from the Indian National Science Academy (INSA) and the Indian Committee of URSI supported NWASS 2010 as a regional event of URSI Commission F.

In recent years, researches in atmospheric sciences have been given a major thrust through various national programmes. The University of Calcutta has a rich tradition of radio and space research pioneered by Late Professor S. K. Mitra. In keeping with the record, S. K. Mitra Centre for Research in Space Environment was set up by the University in the year 2002 as an inter-departmental centre to facilitate and co-ordinate researches in the space environment spanning from the lower to upper atmosphere. A research base has been created in the areas of lower atmospheric chemistry and aerosol studies, radio remote sensing of atmosphere, radio propagation, atmospheric electricity, ionospheric studies and space weather. The centre has also participated in a number of National and International programmes such as, CAWES-India, Megha Tropiques, Ka-band propagation studies, GAGAN, ARFI Project, ISRO Space Science Promotion Scheme and SCINDA.

This National Workshop was planned to aware the students, researchers and faculty members, particularly from eastern and north-eastern part of the country, of the state of the art in space and radio research activities and to encourage them to pursue researches in these areas as their long-term career options. Eminent scientists and academicians from national institutions, ISRO centres and Universities delivered lectures in their fields of expertise. There were also sessions for contributed papers giving preference to young scientists and researchers.

The topics which were covered in the workshop are Radio Propagation and Communication, Space Weather, Remote Sensing, Atmospheric Science, GPS Applications, Radio Astronomy and Planetary Science.

BOOK PUBLISHED FOR URSI RADIOSCIENTISTS

The Principles of Astronomical Telescope Design

by Jingquan Cheng, Astrophysics and Space Science Library, Volume 360, Berlin, Springer 2009, 631 pp.,
ISBN 978-0-387-88790-6; EU 139.00..

The title of this book is a misnomer, to the extent that a coherent development of the fundamental aspects of telescope design is not being presented. Rather, the book offers more of an encyclopedic review of a wide range of facets on the subjects of design, construction, calibration, and operation of telescopes used in all parts of the electromagnetic spectrum. This extends from long radio wavelengths through millimeter, infrared, optical, and ultraviolet wavelengths, all the way to X-ray, gamma ray, and gravitational waves. As such, the book contains a wealth of information. However, for a particular subject – as, for instance, structural design or mirror aberrations – the material is often scattered over several chapters. One needs the extensive and adequate listing of content (six pages) and index (16 pages) to paste the material together, to obtain a complete picture of the problem of interest.

The author is obviously aware of this when he writes in the Preface, "Because many component design principles can be applied to a particular telescope design, readers should reference all relevant chapters and sections when a telescope design project is undertaken." As I experienced during my reading, this is not easy. Although the author mentions that the book is an outgrowth of his lecture notes for postgraduate students, a lecturer will need to have a very complete knowledge of the book's content to effectively guide the students through the material. This obviously was the case with the author as lecturer.

In the first three chapters (220 pages), the author treats fundamental properties, mirror design, and structural design, with control systems of optical telescopes. There follows a chapter on active and adaptive techniques to improve the imaging quality, and a short description of different types of (optical) interferometers. A short chapter on space telescopes completes the treatment of the optical spectrum.

Over the following 160 pages, the attention turns to radio telescopes, with chapters on fundamentals and structural design, as well as a separate chapter on special aspects of millimeter-wavelength telescopes. This field being my major competence, I shall comment here in some more detail, and thereby illustrate the "scattered" treatment of certain problems.

Chapter 6, titled "Fundamentals of Radio Telescopes," begins with a lengthy discussion of receiver sensitivity, the problem of "confusion" by weak background radio sources and the influence of the atmosphere, coupled with telescope-site selection. Only after this unusual introduction are the basic antenna parameters treated. This is one example of the idiosyncratic way the author selects the order of his presentation.

The text contains quite a number of small errors or a lack of precision, which a knowledgeable reader will detect, but which might confuse the student. For instance, Equation (6.13) must be integrated over the entire sphere, not the main-beam solid angle; Equations (6.12a) and (15) are the convolution of the beam and source-brightness distribution, not the received power when pointing at the source. Table 6.3 contains several errors or dubious statements: in line three, the entries "easy-difficult" should be reversed; in line eight, the effect of axial defocus of the feed in the Cassegrain is smaller proportional to m^2, not m; and in the last line, it should be "m times smaller," not larger. It is likely that such errors also occur in the other chapters, but have escaped detection because of my lack of detailed knowledge. These errors decrease the value of the text as a reliable reference for the beginner in the field.

Chapter 7, "Radio Telescope Design," contains extensive discussions of tolerances in reflector-surface precision, positional errors, aperture blocking, and "best fitting" procedures. Again, one finds partial repetitions of material presented earlier in the book without cross-reference: for instance, the treatment of blocking on pages 365-368 and 396-400. The section on structural design is very short, but wind influences and aspects of active control are described in more detail. The chapter closes with a description of radio interferometers, which partially overlaps with earlier discussions.

Radio astronomers, along with electronics and structural engineers, have pushed telescope technology to explore the very shortest radio waves, at frequencies up to 1 THz. The special problems encountered in the realization of telescopes for this regime are discussed in Chapter 8. The critical aspect of thermal effects on the telescope's structure and reflector panels is treated in detail. Carbon-fiber-reinforced plastic has found a wide application in sub-mm-telescopes, due to its superior thermal characteristics. An extensive subsection is devoted to its properties and application. The chapter closes with a highly mathematical discussion of the signal-to-noise aspects of the holographic method of reflector-surface measurement. Unfortunately, the practical application is not discussed.

Chapter 9 deals with infrared, ultraviolet, X-ray, and gamma-ray telescopes, i.e., instruments that are mostly operated from space, while gravitational-wave, cosmic, and dark-matter telescopes are introduced in Chapter 10. A general description of the characteristics of these types of radiation and the methods of detection is followed by short descriptions of the major telescopes in existence or planned. A short chapter with tables of all major telescopes in the world closes the book. Unfortunately, the tables contain quite a number of inaccuracies, but the overall listing is a useful reference.

In summary, this book covers an enormous array of aspects related to telescopes for the entire EM spectrum at a varying depth. There is a lack of connection between the many – mostly, not derived – formulas and their practical application. Also, one misses examples of actual construction and operational facets of the different telescopes. The presentation lacks coherence: often, a certain aspect is treated partially at widely different places in the text. The book will be most useful to readers with an understanding of at least some type of telescope, who want to obtain a feeling for how problems in other wavelength regimes have been attacked. It is more an encyclopedic reference volume than a well-structured introduction to the *Principles of Astronomical Telescope Design*. A better title might have been, "Aspects and Methods of Astronomical Telescope Design."

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International Geophysical Calendar 2011

	S	M	T	W	T	F	S
JANUARY							1
	2	3	4 N	5	6	7	8
	9	10	11 *	12 *	13	14	15
	16	17	18	19 F	20	21	22
	23	24	25	26	27	28	29
	30	31	1	2	3 N	4	5
FEBRUARY	6	7	8 *	9 *	10	11	12
	13	14	15	16	17	18 F	19
	20	21	22	23	24	25	26
	27	28	1	2	3	4 N	5
MARCH	6	7	8 *	9 *	10	11	12
	13	14	15	16	17	18	19 F
	20	21	22	23	24	25	26
	27	28	29	30	31	1	2
APRIL	3 N	4	5 *	6 *	7	8	9
	10	11	12	13	14	15	16
	17	18 F	19	20	21	22	23
	24	25	26	27	28	29	30
MAY	1	2	3 N *	4 *	5	6	7
	8	9	10	11	12	13	14
	15	16	17 F	18	19	20	21
	22	23	24	25	26	27	28
	29	30	31	1 N	2	3	4
JUNE	5	6	7	8	9	10	11
	12	13	14	15 F	16	17	18
	19	20	21	22 *	23 *	24	25
	26	27	28	29	30		
	S	M	T	W	T	F	S

S	M	T	W	T	F	S	
					1 N	2	JULY
3	4	5	6	7	8	9	
10	11	12	13	14	15 F	16	
17	18	19	20	21	22	23	
24	25	26	27 *	28 *	29	30 N	
31	1	2	3	4	5	6	AUGUST
7	8	9	10	11	12	13 F	
14	15	16	17	18	19	20	
21	22	23	24 *	25 *	26	27	
28	29 N	30	31	1	2	3	SEPTEMBER
4	5	6	7	8	9	10	
11	12 F	13	14	15	16	17	
18	19	20	21	22	23	24	
25	26	27 N *	28 *	29	30	1	
2	3	4	5	6	7	8	OCTOBER
9	10	11	12 F	13	14	15	
16	17	18	19	20	21	22	
23	24	25 *	26 N *	27	28	29	
30	31	1	2	3	4	5	NOVEMBER
6	7	8	9	10 F	11	12	
13	14	15	16	17	18	19	
20	21	22	23 *	24 *	25 N	26	
27	28	29	30	1	2	3	DECEMBER
4	5	6	7	8	9	10 F	
11	12	13	14	15	16	17	
18	19	20	21 *	22 *	23	24 N	
25	26	27	28	29	30	31	2012
1	2	3	4	5	6	7	JANUARY
8	9 F	10	11	12	13	14	
15	16	17	18 *	19 *	20	21	
22	23 N	24	25	26	27	28	
29	30	31					
S	M	T	W	T	F	S	

N NEW MOON F FULL MOON

11 Regular World Day (RWD)

12 Priority Regular World Day (PRWD)

9 Quarterly World Day (QWD) also a PRWD and RWD

5 Regular Geophysical Day (RGD)

7 8 World Geophysical Interval (WGI)

+ Incoherent Scatter Coordinated Observation Day

4 Solar Eclipses: **Jan 4, Jun 1, Jul 1, Nov 25 (partial)**

All solar eclipses in 2011 are partial eclipses.

27 28 Airglow and Aurora Period

11* Dark Moon Geophysical Day (DMGD)

This Calendar continues the series begun for the IGY years 1957-58, and is issued annually to recommend dates for solar and geophysical observations which cannot be carried out continuously. Thus, the amount of observational data in existence tends to be larger on Calendar days. The recommendations on data reduction and especially the flow of data to World Data Centers (WDCs) in many instances emphasize Calendar days. The Calendar is prepared by the International Space Environment Service (ISES) with the advice of spokesmen for the various scientific disciplines.

The Solar Eclipses are:
2011 has no full or annular eclipses; it does have four partial eclipses.

a.) 4 January 2011, partial solar eclipse, throughout Europe, northern Africa, east to India; max 86% in Scandinavia; near 75% in London, near 73% Paris, 58% Madrid, 55% Cairo, 10% in Karachi.
b.) 1 June 2011, partial solar eclipse, throughout N. Alaska, N. Canada, Greenland; max 60% N. Siberia near Finland; 46% Reykjavik, 40% Greenland; 25% Prince Edward Island, 12% St Johns, 9% St. Pierre et Miquelon; just NW of Halifax and Fredricton, 3% at Fairbanks, Alaska.
c.) 1 July 2011, partial solar eclipse, visible only in the ocean off Antarctica.
d.) 25 November 2011, partial solar eclipse, in Antarctica, New Zealand, barely Cape Town, South Africa; max 90% in ocean near Antarctica south of S. America; 24% at Nelson (N.Z.), 28% Christchurch; 30% Invercargill, 6% Hobart (Tasmania), 10% Cape Town (South Africa).

Information from Jay M. Pasachoff, Williams College (Williamstown, MA), Chair, International Astronomical Union's WG on Eclipses (http://www.eclipses.info), based on work by Fred Espenak, NASA GSFC and provided as a Google Map by Xavier Jubier.

Meteor Shower
(selected by P. Jenniskens, SETI Institute, MountainView, CA, pjenniskens@mail.arc.nasa.gov) include important visual showers and also unusual showers observable mainly by radio and radar techniques. The dates are given in Note 1 under the Calendar. See extended text for more details.

Definitions
Time = Universal Time (UT);
Regular Geophysical Days (RGD) = each Wednesday;
Regular World Days (RWD) = Tuesday, Wednesday and Thursday near the middle of the month (see calendar);
Priority Regular World Days (PRWD) = the Wednesday RWD;
Quarterly World Days (QWD) = PRWD in the WGI;
World Geophysical Intervals (WGI) = 14 consecutive days each season (see calendar);
ALERTS = occurrence of unusual solar or geophysical conditions, broadcast once daily soon after 0400 UT;
STRATWARM = stratospheric warmings;
Retrospective World Intervals (RWI) = MONSEE study intervals

For more detailed explanations of the definitions, please visit http://www.ngdc.noaa.gov/stp/SOLAR/IGCwebpage5.html or contact H. Coffey or http://www.ises-spaceweather.org/.

Priority recommended programs for measurements not made continuously (in addition to unusual ALERT periods):

Aurora and Airglow — Observation periods are New Moon periods, especially the 7 day intervals on the calendar;
Atmospheric Electricity — Observation periods are the RGD each Wednesday, beginning on 5 January 2011 at 0000 UT, 12 January at 0600 UT, 19 January at 1200 UT, 26 January at 1800 UT, etc. Minimum program is PRWDs.

The International Space Environment Service (ISES) is a permanent scientific service of the International Union of Radio Science (URSI), with the participation of the International Astronomical Union (IAU) and the International Union of Geodesy and Geophysics (IUGG). ISES adheres to the Federation of Astronomical and Geophysical Data Analysis Services (FAGS) of the International Council for Science (ICSU). The ISES coordinates the international aspects of the world days program and rapid data interchange.

This Calendar for 2011 has been drawn up by H.E. Coffey, of the ISES Steering Committee, in association with spokesmen for the various scientific disciplines in SCOSTEP, IAGA, URSI and other ICSU organizations. Similar Calendars are issued annually beginning with the IGY, 1957-58, and are published in various widely available scientific publications. PDF versions are available online at ftp://ftp.ngdc.noaa.gov/STP/SOLAR_DATA/IGC_CALENDAR. The calendar is published for the International Council for Science and with financial assistance of UNESCO.

Copies of earlier years' calendars are available upon request to either ISES Director, Dr. David Boteler, Geomagnetic Laboratory, Natural Resources Canada, 7 Observatory Crescent, Ottawa, Ontario, Canada, K1A 0Y3, FAX (613)824-9803, e-mail dboteler@NRCan.gc.ca, or contact ISES Secretary for World Days, Ms. Helen Coffey, e-mail hecoffey799 at aol.com. Beginning with the 2008 Calendar, all calendars are available only in digital format.

Calendar information is available on-line at http://www.ises-spaceweather.org/. The International Geophysical Calendar and descriptive text is also available online at http://www.ngdc.noaa.gov/stp/solar/onlinepubl.html under International Geophysical Calendar.

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Note: an alphabetical index of names, with coordinates and page references, is given on pages 75-90.

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